Continuous-wave moving target radar using two microwave transmitters.

Author:
Egan, Thomas Michael

Publication Date:
1972

DOI:
https://doi.org/10.26190/unsworks/5177

CONTINUOUS-WAVE MOVING TARGET RADAR

USING

TWO MICROWAVE TRANSMITTERS

BY

THOMAS MICHAEL EGAN

SUBMITTED AS A THESIS

FOR FULFILMENT OF THE DEGREE

OF

MASTER OF SCIENCE

IN THE SCHOOL

OF

APPLIED PHYSICS AND OPTOMETRY

OF

THE UNIVERSITY OF NEW SOUTH WALES

JUNE 1972

The candidate, hereby declares that none of the work presented in this thesis has been submitted to any other University or Institution for a higher degree.

ABSTRACT

A short range CW Doppler radar is described which is capable of either simultaneous measurement of the velocity and the range (and hence the imminence time) of the target or velocity measurement with clutter cancellation. In the latter case the range could be estimated by the amplitude of the Doppler signal.

This thesis reports the work carried out to make, using two separate oscillators, such a CW Doppler radar with a microwave spectrum which has two frequencies f and $f + \Delta f$, such that f is approximately 10 GHz and Δf is approximately 1 MHz, by using an automatic difference frequency control circuit to maintain the value of Δf constant to within $\pm$ 5%.

The suitability of solid-state oscillators for this purpose is discussed especially with regard to their frequency stability and FM noise figures. Various locking phenomena are described and it is indicated that the frequencies of the two solid-state oscillators were able to be controlled so that their frequency difference was less than the locking bandwidth for typical isolations of -40 dB.

A design for a digital frequency discriminator, which has a sampling rate of approximately 8 kHz, is given together with methods of controlling the frequency of a solid-state microwave oscillator. Varying the capacitance of a varactor diode coupled to the frequency determining cavity was chosen for experimentation.

Means of feeding one antenna in order to provide the required isolation of the two microwave oscillators, and the correct coupling of each oscillator to its respective mixer (while at the same time giving a range of up to 500 feet), are reported. Evidence is presented to show that control of this isolation and coupling and not the difference frequency control is the greater technical problem.

Results obtained, with the constructed radar, show that the expected phase difference variation between the two separate Doppler signals (one for each transmitted frequency) as the range of the target changed, was linear and correct only when the above isolation and coupling were sufficient.

A suggestion is made to construct a short range CW radar with a three frequency microwave spectrum. This spectrum would be produced by frequency modulating a single microwave oscillator with a three-step staircase waveform. The three Doppler signals, produced (one for each transmitted microwave frequency) by synchronously switching the mixer output, would be input, in pairs, to two differential amplifiers in order to give clutter cancellation, while the phase/time difference between the outputs of these amplifiers would give range/imminence time information.

ACKNOWLEDGEMENTS

The author wishes to thank the following:

Professor C.J. MILNER, for many interesting and informative discussions about the ideas involved in clutter cancellation and/or range measurement using a two/three frequency microwave spectrum; N. GILLIES, for some conversations about operational amplifiers and control circuits; J. BOLIN, for construction of the varactor diode cavity and the Gunn diode cooling fins; R. GRAVES, for wiring up and testing the video-amplifier; M. PISCH, for taking and developing the negatives of Fig. 10 to Fig. 30, and Fig. 36; P. MARTIN, for printing and mounting the photographs; R.D. MARTIN, for drawing Fig. 1 to Fig. 9 and Fig. 31 to Fig. 35; and R. MARTIN, for typing the original manuscript.

<u>SYMBOL DEFINITION</u>

r	=	the radius vector of a moving target (m).
v	=	the line-of-sight, or radial velocity of the target (m/sec).
T	=	the imminence time of a moving target (sec).
	=	r/v (sec).
Ω	=	the angular velocity of the target (radian/sec).
f_D	=	a Doppler frequency (Hz).
T_1	=	$1/f_D$, the periodic time of a Doppler signal (sec).
f	=	a microwave frequency (Hz).
c	=	the velocity of microwave radiation (m/sec).
f_1	=	the microwave frequency of the Gunn diode oscillator 1 (Hz).
f_2	=	the microwave frequency of the Gunn diode oscillator 2 (Hz).
Δf	=	$f_1 - f_2$ (Hz)
	=	the frequency difference between the two microwave oscillators, 1 and 2, with $f_1 > f_2$.
f_{D1}	=	the Doppler frequency associated with the microwave frequency f_1 (Hz).
f_{D2}	=	the Doppler frequency associated with the microwave frequency f_2 (Hz).
$\Delta\Phi$	=	the phase-difference between the Doppler signals f_{D1} and f_{D2} (radian).

τ = the time-difference between the respective zero-crossings of the two Doppler signals with frequencies f_{D1} and f_{D2} (sec).

ϕ = the instantaneous phase-difference between the two microwave oscillators (radian).

P_1 = the locking (injected) oscillator power (mW).

P_2 = the locked oscillator power (mW).

Q_L = the loaded Q of the system.

ω_o = the nominal angular frequency of the two oscillators (radian/sec).

$\Delta\omega_o$ = $2\pi\Delta f$ (radian/sec).

η = $1/Q_L$, figure of merit of the system.

$\partial f/\partial V$ = change in operating frequency of a Gunn diode with bias voltage variation (MHz/V).

$\partial f/\partial T$ = change in operating frequency of a Gunn diode with temperature variation of the diode itself (MHz/^{o}C).

C_o = the varactor diode capacitance at 0 volts reverse bias (pF).

C_v = the varactor diode capacitance at V volts reverse bias (pF).

V_{CL} = the control voltage applied to the varactor diode coupled to the cavity of one of the Gunn diode oscillators (Volt).

CONTENTS

CHAPTER ONE

SHORT RANGE CW RADAR

1.1 INTRODUCTION

The utility of CW radar systems is well described by HANSEN[22], p132, "... little information may be obtained with ease, where a lot is impossible to obtain. Such a situation arises in the presence of very severe clutter, where pulse systems, even with MTI equipment, may fail to give any information." This information about a single target may be listed as various radar parameters of the target, such as, 1) angular velocity (Ω), 2) radial (or line-of-sight) velocity (v), 3) range (r) and 4) imminence time (T). For more than one target, the value of the parameters measured would be assigned to that target which gave the most intense reflection.

A measurement of Ω can be obtained with the use of either a rotating antenna or a phased array antenna. The Doppler frequency will be proportional to the radial velocity, v. Some form of modulation of the transmitted microwave energy is usually required for a range measurement; amplitude modulation and frequency modulation are the most frequently used. Range and radial velocity, when measured, give the imminence time thus,

$$T = r/v \text{ (sec)} \tag{1}$$

A precise knowledge of this imminence time is very valuable in systems designed to avoid collisions with oncoming targets or designed to detect and collide with the oncoming target (i.e. "homing" devices). A short-range CW Radar capable of measuring r and v simultaneously would seem to be well suited to the above two applications.

1.2 RADAR PARAMETERS

1.2.1 Radial Velocity (v)

A measure of the Doppler frequency (f_D) will give an estimate of the radial velocity, such that

$$f_D = \frac{2fv}{c} \quad (Hz) \tag{2}$$

For f approximately equal to 10^{10} Hz and c approximately equal to 3×10^8 m/s we have typical Doppler frequencies for motor vehicles with $V = 60$ m/s

$$f_D = 3.6 \times 10^3 \text{ (Hz)} \tag{3}$$

and for aircraft with $v = 600$ m/s

$$f_D = 3.6 \times 10^4 \text{ (Hz)} \tag{4}$$

Equations (3) and (4) would give an idea of the upper cut-off frequency (3 dB point) required of the Doppler channel amplifier. The lower cut-off frequency (3 dB point) for the Doppler channel amplifier would be fixed by a consideration of the lowest Doppler frequency expected.

The "Doppler" frequencies to be considered are not only those caused by relative motion between the target and the radar (including those due to a rotating antenna) but also those due to microphony caused by either (a) vibration of the microwave oscillator and cavity, or (b) vibration resulting in a relative motion between the antenna and nearby reflectors or (c) reflections from the ground (ground clutter). (This interference from microphony can, of course, be at any frequency depending on the magnitude of the relative velocities concerned). For a system using an intermediate frequency, equations (3) and (4) would give the half-bandwidth expected of the intermediate frequency amplifiers. A rotating antenna or phased-array antenna, used to estimate Ω, would introduce low frequency amplitude modulation on the Doppler signal. The frequencies introduced to the Doppler frequency spectrum depend on the beam width and angular velocity of the beam sweep. Typical values would be of the order of 50 Hz.

1.2.2 Range (r)

A common characteristic of CW radars that are capable of measuring range seems to be that their respective microwave frequency spectra contain more than one frequency. The relative phases between the transmitted and reflected microwave frequencies can be processed in various ways to obtain a measurement of range. (SKOLNIK[43,44]).

A simple microwave frequency spectrum is one which contains two frequencies, spaced Δf (Hz) apart. For $\Delta f \ll f$, where f = the microwave frequency, each microwave frequency will have associated with it a Doppler signal (see equation (2)) for moving targets. BOYER[8], HANSEN[22] and SKOLNIK[43,44] point out the following, a) the respective Doppler signals have frequencies which are approximately equal and calculated by equation (2), b) the respective Doppler signals have a phase-difference $\Delta\Phi$ (Radians) given by

$$\Delta\Phi \propto \Delta f . r, \tag{5}$$

(c) from equation (5) it is to be noted that $\Delta\Phi = 0$ for $r = 0$ (for $\Delta f = 0.5$ MHz, $\Delta\Phi = \pi$ radians when $r = 150$ m), and d) the relative timing (τ) of the positive going zero-crossings of the two Doppler signals for an approaching target will be reversed for receding targets, i.e. there is a reversal of the phase-difference between the two Doppler signals; see Fig.2.

1.2.3 Imminence Time

For a two-frequency microwave spectrum, where Δf remains constant with respect to time,
we have from (5)

$$\Delta\Phi \propto r \tag{6}$$

and from (2)

$$f_D \propto v \tag{7}$$

Hence
$$\frac{\Delta\Phi}{f_D} \propto \frac{r}{v} \propto T \tag{8}$$

where T = the imminence time (sec).

Also from Fig.2 we have

$$\frac{\Delta\Phi}{2\pi} = \frac{\tau}{T_1} \tag{9}$$

where $$f_D = \frac{1}{T_1} \tag{10}$$

i.e. T_1 = the periodic time of the Doppler frequency.

From (8) and (9) we obtain

$$\frac{\Delta\Phi}{f_D} \propto T \propto \tau \tag{11}$$

Hence we have, from (11)

$$\tau \propto T \tag{12}$$

The constant of proportionality is given by

$$\tau = \frac{\Delta f}{f}\ T \text{ (sec)} \tag{13}$$

e.g. T = 5 seconds, Δf = 0.5 MHz, f = 10^{10} Hz gives τ = 250 µsec, or (50 µsec)/(second imminence time). (See MILNER[39], p9).

The relative phase of the two Doppler signals is shown in Fig.2 for approaching targets. Assuming that the target will not suddenly reverse its direction, an increase or decrease of the parameter τ should indicate a receding or approaching target.

1.3 SUMMARY

The four main points (Section 1.2.2) form the basis for the work to be described in this thesis. In particular the construction of and results obtained from a two solid-state oscillator CW radar system will be discussed, together with a description of the technical problems encountered. The concluding chapter will contain comments and tentative explanations of the various observed phenomena and a suggestion for further work.

CHAPTER TWO

THE PRODUCTION OF A TWO FREQUENCY MICROWAVE SPECTRUM

2.1 INTRODUCTION

As indicated in 1.2.2, for phase differences between the two Doppler signals of 2π at $r = 150$ m, we have $\Delta f = 1.0$ MHz. There would seem to be four major methods of obtaining two microwave frequencies spaced approximately 1.0 MHz apart: 1) frequency multiplication of a base frequency of 1.0 MHz, and tune to two neighbouring harmonics in the microwave region (e.g. X-band), 2) amplitude modulation of a single microwave frequency, (the amplitude modulation frequency being equal to 1.0 MHz), and tune to the carrier and one of the sidebands, 3) frequency diplexing, by switching a single microwave oscillator between two microwave frequencies with a spacing of 1.0 MHz, and 4) automatic frequency control, using two microwave oscillators, to keep the two frequencies 1.0 MHz apart.

2.2 FREQUENCY MULTIPLICATION METHOD

This method has four essential requirements; 1) a base frequency oscillator with a frequency approximately 1.0 MHz, which is to be stable to something like 1%, for a similar stability in Δf. This stability is to hold for ambient temperature and humidity variations as well as mechanical vibrations, 2) a frequency multiplier chain capable of reaching the microwave region of the spectrum while still maintaining good phase stability and without introducing excessive noise, 3) a microwave power output with sufficient range, for Doppler frequency signals, of up to 200 m. Some form of phase-locking might also be applicable to this method to give the required stability and reduce the noise in the microwave power output, 4) a method of tuning to the two chosen harmonics so that the two crystal mixers will be sufficiently selective to their respective microwave local oscillations and thus produce the relevant Doppler signal for each transmitted microwave frequency. These tuning circuits are to have a sufficient Q to exclude the other harmonics and yet include the expected Doppler frequencies. This value of Q is also to be maintained for ambient temperature, and humidity variations and mechanical vibrations.

These requirements would seem to make this method not a very promising one for construction of a two frequency Doppler CW radar. The two major difficulties would be in the multiplier being able to give sufficient microwave power output with low noise suitable for a 200 m range Doppler radar, and in the ability of the tuned circuits to be able to tune to the required harmonics and attenuate the other harmonics sufficiently to avoid cross-products in the mixer output.

2.3 AMPLITUDE MODULATION METHOD

Amplitude modulation, at a frequency Δf Hz, of a microwave frequency will give two side-bands at $\pm\Delta f$.[36] The relative energies of the carrier and either side-band will be fixed by the modulation index. Tuning to the carrier and one side-band would waste the energy of the other side-band. Tuning to each side-band would waste the energy of the carrier. The carrier and the side-bands would not in general have the same energy, hence the respective transmitted energies would be unequal. The oscillator would therefore be required to have more power available than the maximum range power for the carrier alone. It is understood that the tuning of two cavities to microwave frequencies Δf apart would not be difficult but to hold this tuning for variations in temperature and humidity as well as for mechanical vibrations would present some technical problems. Some monitoring of Δf would be required to ensure that the correct value of Δf was being maintained. This would suggest the need for some form of control, either of the amplitude modulation frequency, or the tuning of the cavities. Again, as in 2.2 above, it is essential that the two mixers are selective to their respective LO, so that each will produce the correct Doppler signal relevant to each transmitted microwave frequency.

An interesting approach would be to use a balanced modulator and suppress the carrier, thus making all the energy available in the two side-bands, which would also be of equal amplitude.

2.4 DIPLEXING METHOD

The diplexing method is fully described, with equations, in BOYER'S paper.[8] The oscillator chosen was a reflex klystron whose microwave frequency was switched at a rate of 100 kHz between two frequencies spaced Δf apart. The square wave modulation voltage was applied to the repeller, with an amplitude of the order of 0.5V/MHz difference frequency. The two Doppler signals are also synchronously switched to two separate, but identical, Doppler frequency amplifiers. Low pass filtering, to remove the 100 kHz frequency component, ensures that each Doppler frequency amplifier passes a Doppler signal relative to its respective microwave frequency. The production of Δf by this method is straight forward. Synchronous switching presents problems, so does a frequency/phase matching of the Doppler frequency amplifiers. A Gunn diode, mounted in a waveguide or co-axial cavity would be more suitable for a mobile radar, since its power supply requirements are less complex than those of a reflex klystron. Frequency switching could be achieved by a varactor diode coupled to the waveguide or co-axial cavity. A voltage square wave applied to the varactor diode will change the frequency of oscillation by Δf Hz. (The Gunn diode is also much less microphonic than a reflex klystron). CLEGG and CROMPTON[11] use a "J2" method to cancel the effect of strong local reflections and the relative mechanical movement of microwave oscillator and antenna, both of which produce large amplitude signals of similar frequency to any desired Doppler signals. From equation (5) we saw, (in Chapter 1), that at zero-range the phase difference between the two Doppler signals would be zero. Hence having the Doppler signals as the two inputs to a differential amplifier would give considerable cancellation of the strong local-reflection and movement-of-antenna type Doppler signals. The inputs to the differential amplifier would need to be of a similar amplitude to give good cancellation. It would also seem that for this application a square wave frequency modulation is not critical. (While for BOYER's method[8], it is).

The "J2" system is an example of the choice of a frequency modulation waveform that gives small amplitude Doppler signals from local reflections, i.e. the phases of the various Doppler signals, associated with each frequency in the radio-frequency spectrum, are such as to cancel for small r. The "diplexing" type of frequency modulation produces a radio-frequency spectrum that gives two Doppler signals which are in phase for r = 0. In general, for local reflection cancellation, a frequency modulation waveform should be chosen that gives a known and useable phase relationship for the Doppler signals as r approaches 0.

2.5 AUTOMATIC DIFFERENCE FREQUENCY CONTROL METHOD

Various methods, using AFC, of stabilising the frequency of a microwave oscillator, in particular the klystron oscillator, are discussed by SMITH[45], HARVEY[23] and OWSTON[34]. SMITH uses cavities tuned to frequencies just above and just below the desired microwave frequency. (A Travis type AFC). OWSTON uses amplitude modulation and a phase sensitive detector to obtain the required stability. HARVEY, p940, gives a good summary of the various methods available for stabilising klystrons. Most of the methods use cavities tuned to the various microwave frequencies; one of the best DC types being that due to POUND (173, 174, ref. 23). An AC POUND type stabilising circuit is described which uses an amplitude modulation of the microwave energy at a frequency of 45 MHz. These references 45, 23, 34 would suggest that it is possible to have cavities tuned to microwave frequencies spaced Δf apart. The values of Δf discussed is usually a few tens of MHz rather than the one MHz required for a measurement of range as proposed by BOYER[8]. HARVEY, p945, notes the difficulty of tuning two cavities to give the required overlap at a definite frequency.

To produce a two-frequency microwave spectrum, two microwave oscillators would be required one of which is frequency stabilised such that the difference frequency between the oscillators is controlled at the preset value, typically 1 MHz. Two microwave oscillators with frequencies as close as 1 MHz will, in general, be liable to frequency locking. Any control circuit used should, therefore, be capable of keeping the oscillators from locking; to do this a fast response time is required. Too fast a response time will, of course, produce "hunting"; some form of integrator is therefore necessary.

2.6 SUMMARY

Amplitude modulation, diplexing and automatic difference frequency control seem the most promising methods for producing a two-frequency microwave spectrum. BOYER[8] has used the diplexing method with success, with a klystron oscillator. Of the two remaining methods it was felt that the control method was in essence the simpler; it was, therefore, chosen for experimentation.

CHAPTER THREE

LOCKING PHENOMENA IN OSCILLATORS

3.1 INTRODUCTION

Two oscillators, with frequencies Δf Hz apart, give rise to locking phenomena in which Δf approaches zero. The basic equation for this effect is given by ADLER[1] thus

$$\frac{d\phi}{dt} = -\left[\frac{P_1}{P_2}\right]^{\frac{1}{2}} \frac{\omega_o}{2Q_L} \operatorname{Sin}(\phi) + \Delta\omega_o \tag{14}$$

where ϕ = instantaneous phase difference between the two oscillators

P_1 = locking power (Injected)

P_2 = locked power

Q_L = loaded Q of the system

ω_o = nominal angular frequency of the oscillators (ω_o usually $\gg \Delta\omega_o$)

$\Delta\omega_o$ = $2\pi\Delta f$

The condition for synchronisation (i.e. that $d\phi/dt$ should be zero) is also, from (14)

$$\left[\frac{P_1}{P_2}\right]^{\frac{1}{2}} > 2Q_L\left|\frac{\Delta\omega_o}{\omega_o}\right| \tag{15}$$

Rewriting (14) we have

$$\frac{d\phi}{dt} = -B \operatorname{Sin}(\phi) + \Delta\omega_o \tag{16}$$

where $B = \left[\frac{P_1}{P_2}\right]^{\frac{1}{2}} \frac{\omega_o}{2Q_L}$ (17)

From equation (16), as B approaches zero, (i.e. as the ratio of the locking power to the locked power approaches zero), $d\phi/dt$ will approach $\Delta\omega_o$. Under these conditions, i.e. a large amount of isolation between the two oscillators, the rate of change of phase-difference between the two oscillators approaches the difference frequency for the two oscillators. Hence the output of a crystal mixer, receiving the isolated output from both oscillators, will be a sinusoidal waveform at the difference frequency Δf. As B increases, the output waveform of the crystal mixer, which is proportional to $d\phi/dt$, becomes significantly non-sinusoidal until the synchronisation condition, given by equation (15), is reached. (See ADLER[1], COUCH[13], and HUNTOON[28]). When the two oscillators are synchronised, the mixer output will consist of a constant term added to the FM noise of the oscillators.

3.2 ISOLATION VERSUS LOCKING BANDWIDTH

For $\Delta f = 0.5 \times 10^6$ Hz, and $f = 10^{10}$ Hz we have from equation (15)

$$\left[\frac{P_1}{P_2}\right]^{\frac{1}{2}} = Q_L \times 10^{-4} \tag{18}$$

Typical loaded Q_L values are 100, for co-axial cavities (AITCHISON[3], FANK[18] and FITZSIMMONS[19]), while typical Q_L values for waveguide cavities are 500 (FANK[18] and HOBSON[26]).

Taking $Q_L = 100$ we therefore have

$$\left[\frac{P_1}{P_2}\right]^{\frac{1}{2}} = 10^{-2}$$

and

$$\left[\frac{P_1}{P_2}\right] = 10^{-4} \text{ , i.e. 40 dB} \tag{19}$$

This sort of isolation is confirmed by HOLLIDAY[27], HOBSON[26], DAY[15] and FITZSIMMONS[19]. HOBSON[25] suggests 10 dB of attenuation near the cavity output to avoid phase locking, plus the use of a hybrid-tee to give sufficient isolation. JOCHEN[31] puts forward the idea that it is the internal resistance R_I (of the Impatt or Gunn diode) which controls the locking bandwidth rather than the loaded Q_L. BOTT[7] and HOLLIDAY[27] obtain a figure of merit, η, such that

$$\eta = \frac{2\Delta\omega_o}{\omega_o}\left[\frac{P_2}{P_1}\right]^{\frac{1}{2}} \tag{20}$$

Comparing with equation (15) we have

$$\eta = \frac{1}{Q_L} \tag{21}$$

Values of η of .0016 to .05 are given. HOLLIDAY also uses an isolator, circulator and precision attenuators to obtain the required isolation, and indicates how η is critically dependent on the value of the various microwave parameters of the circuit. COUCH[13] and BACK[5] make the point that the locking bandwidth of two coupled oscillators is a function of the electrical spacing between the oscillators as well as of their relative powers.

3.3 SOLID-STATE OSCILLATORS

JOCHEN[31] confirms ADLER's theory that the locking bandwidth is proportional to $(P)^{\frac{1}{2}}$ where P = injected power for Impatt and Gunn diodes. This reference also suggests that a considerable increase in the locking bandwidth can be obtained when "... the oscillator diode itself is matched to the waveguide by means of a radial transmission line inside the resonator instead of a thin voltage supplying cylindrical post." A strong warning is given to avoid local power reflections which tend to reduce the locking bandwidth. The last paragraph is of some interest for the present study; it is stated "there remains a strong discrepancy of locking bandwidth with loaded Q-factor and the locked amplitude response ... a detailed equivalent circuit of synchronised microwave oscillators is unknown."

DAY[15] maintains that ADLER's equations seem to apply to Gallium Arsenide but are also dependent on the nature of the Gallium Arsenide samples. Here again the equivalent circuit parameters for a reduced height waveguide, are not known hence the quantitative locking characteristics are not predictable.

COUCH[13] gives a very good description of a phase locked loop and gives the locking properties of two oscillators in terms of their locking bandwidth and locking signatures (See also HUNTOON[28]). For $P_2 \gg P_1$ the amplitude modulation of the oscillator due to synchronisation phenomena is negligible. Experimental details given could be used, to good effect, to study the phase between two Gunn diodes within the locking bandwidth.

BACK[5] indicates that reverse locking can occur due to leakage in a real circulator, and uses ADLER's equations, when the locking signals are small, to obtain a pair of coupled ADLER differential equations for the two oscillators. (A transistor and a klystron oscillator, with f = 3 GHz).

3.4 SUMMARY

ADLER's equations give some idea of the design features required for the use of two oscillators, with sufficient coupling to produce a difference frequency in the region of 1 MHz. Most of the references mentioned agree that the degree of isolation required at this value of difference frequency is of the order 30 to 40 dB. Loaded Q_L and critical matching (including considerations of electrical length between the oscillators) are the most important parameters to optimise. Qualitative agreement with ADLER's equations is obtained for solid-state oscillators such as Impatt and Gunn diodes.

CHAPTER FOUR

SOLID-STATE MICROWAVE OSCILLATORS

4.1 INTRODUCTION

Solid-state microwave sources have two main features to be considered in relation to their use in a CW Doppler radar; these features are 1) Input/Output power characteristics and 2) Frequency Modulation (FM) Noise. When used in conjunction with a cavity, to produce a definite microwave frequency output, two more features are of interest, 1) Output frequency variation with temperature and bias supply voltage and 2) correct and suitable cavity design. The first two features concern properties of the solid-state diodes themselves, while the second two concern properties of the solid-state diodes together with a frequency determining cavity.

4.2 INPUT/OUTPUT POWER CHARACTERISTICS

A literature survey indicates that, of the solid-state microwave oscillators available, the Gunn diode is the most studied and documented. The Impatt diode will be the next on the list. Typical input power requirements for Gunn diodes are 9V at 400 mA for a 50 mW microwave power output, (Mullard 820 CXY/A) at a frequency of 8 to 12 GHz. While a silicon Impatt diode (Hewlett-Packard 5082) would require 90V at 30 mA for a 100 mW microwave power output at a frequency of 8 to 12 GHz. The lower bias supply voltage for the Gunn diode can be supplied by a lead accumulator while the higher bias required by an Impatt diode will have to be provided by a more elaborate power supply. It is to be noted that a single frequency continuous wave (CW) microwave power output is usually obtained by coupling the Gunn or Impatt diode to a waveguide or co-axial cavity tuned to the required frequency. De-coupling of the bias supply, using a low inductance capacitor, is recommended to avoid low frequency oscillations of the circuit.

FANK[18] and FITZSIMMONS[19] give a good general description of the theory of bulk effect devices. Fig.3 of ref.19 gives the relative power output for various solid-state diodes. In particular FANK points out that the transit time mode is not used because of its low efficiency and poor pushing and pulling factors. The domain mode of oscillation is used most for microwave power output less than 100 mW, at frequencies from 4 to 60 GHz. FANK also gives a list of the different modes of oscillation of bulk effect diodes together with their relative microwave power outputs. FANK makes the point that the oscillators (i.e. diode plus cavity) should be decoupled 2 to 3 dB from the maximum power output, for stability of the power output. Also to maintain this stability in the power output, "a maximum bias voltage of 3 times the threshold is necessary to limit saturation of the diode" (see also BYBOKAS[9]). BIRD[6] gives a typical value of 20 mW power output at X-band, with 2% efficiency, using a $\lambda/2$ co-axial cavity with an external loaded Q of 50. HOBSON[26] outlines a typical bias circuit.

4.3 FREQUENCY STABILITY

4.3.1 Frequency Variation with Temperature

HOBSON[25] gives 2 MHz/°C variation of frequency with temperature for a Gunn diode. A correlation between this temperature dependence of the oscillation frequency and the frequency modulation of the Gunn diode through the bias current variations is suggested. The bias current variations cause a temperature variation in the Gunn diode encapsulation itself. HOLLIDAY[27] and BOTT[7] also maintain that a change in bias voltage causes the diode's temperature to change. A 0.5 to 1 MHz/°C variation could be easily obtained without taking too many precautions to overcome the frequency variation with temperature (see also ALBRECHT[4], BYBOKAS[9], DAY[15], KOOI[33], SELING[41] and WARNER[48]). A constant voltage bias supply is therefore, a basic frequency stability requirement. ELLIS[17] points out the effect that the loaded Q of the circuit has on the frequency stability; the higher the loaded Q, the less dependent is the frequency variation on any small change in the value of the bias supply voltage.

The type of cavity, used with the Gunn diode, has a dramatic effect on the frequency stability of the microwave output. COURT[14] suggests the use of a $2\lambda_g$ cavity to increase the frequency stability of the Gunn diode, while FANK[18] obtains 50 kHz/°C and with a multiple cavity design quotes a variation of 10 to 15 kHz/°C. SELING[41] improves a -0.8 MHz/°C with a 40 minute warm up period uncompensated diode, to a -0.03 MHz/°C with a warm up period of 4 minutes compensated diode. This is also one of the few references that mentions the necessity of a warm up period to obtain the maximum stability of the microwave output. BIRD[6] using n^+ contacts, which considerably lower the contact resistance of the Gunn diode, obtains 20 mW of microwave power from a low Q Invar co-axial cavity, decoupled from the maximum power output by 6 dB. The frequency variation with change in temperature is -70 kHz/°C. (Silver-tin contacts used in Gunn diodes usually give something like 1 MHz/°C variation in output frequency). Note the mention of decoupling the diodes to improve the frequency stability. BYBOKAS[9] holds that the decoupling increases the loaded Q; 2 dB of decoupling should reduce the frequency drift by 25 to 50%. An Invar short circuit, in the waveguide cavity, should also increase the frequency stability 50%. ITO[29] makes the point "... providing proper protection from humidity greatly improves the frequency versus temperature coefficient." -7×10^{-7}/°C using an Invar cavity with a brass short circuit was achieved. KOOI[33] using dielectric loading in the cavity obtained -0.6 MHz/°C to +0.054 MHz/°C; the point is also made that the cavity is still stable for use with other diodes. FANK[18] maintains that to go lower than 50 kHz/°C usually requires an additional stable cavity inserted in series with the standard low Q cavity. "... with simple multiple-cavity designs, temperature coefficients of 10 to 15 kHz/°C can be accomplished."

4.3.2 Frequency Variation with Bias Voltage

WARNER[48] gives a typical case for a Gunn diode that for a bias voltage variation from the threshold at 4.8V to a voltage of 7.0V, the frequency variation observed was in the region of 85 MHz. SELING[41] quotes 9 MHz/V for a DC 1414A (Monsanto) 10 mW Gunn diode. ELLIS[17] describes a stripline Gunn diode using the quenched-domain mode, where the frequency of oscillation is determined by the resonance properties of the circuit. End-coupling gives a -50 MHz/V while side-coupling gives a -26 MHz/V variation of oscillation frequency with bias supply voltage. A 1 kHz stability in the output frequency would, therefore, require a voltage stability of something like 50 μV in the bias supply to the Gunn diode. DeSa[16] postulates a direct relationship between $\frac{\partial f}{\partial v}$ and $\frac{\partial f}{\partial T}$, with typical values (at 20^{o}C) of +10 MHz/V and -0.98 MHz/^{o}C respectively.

4.4 FREQUENCY MODULATION NOISE

The frequency modulation noise measured for (solid-state) diodes compares favourably with that measured for klystrons of similar microwave power output, (FITZSIMMONS[19] and WARNER[48] (Fig.4)), at least 200 kHz from the carrier frequency.

COURT[14] indicates that the frequency modulation noise is due to capacitance variation of the diode itself. Waveguide resonators give better frequency modulation noise figures than co-axial cavities. The improvement is due to the increase in the characteristic impedance of the waveguide cavity. Ref.37 distinguishes for Gunn diodes between the transit time mode, where the frequency modulation noise would possibly be due to electron scattering and fluctuations in carrier concentration, and the cavity controlled high-field domain oscillations where the frequency modulation noise would possibly be caused by susceptance fluctuations. The higher the Q, the lower the frequency modulation noise. (See also ITO[29], p892). CLUNIE[12] and SJOLUND[42] discuss the frequency modulation noise in Impatt diodes, while BYBOKAS[9] maintains that a Gunn diode mounted in a flange, would have a 10 dB improvement in frequency modulation noise figure than the Impatt diode. A certain amount of decoupling of the cavity also improves both the amplitude and frequency modulation noise figure for the Gunn diode.

WARNER[48] notes that the peak to peak ripple on the bias voltage supply should be less than 10 μV for the solid state oscillator to approach the frequency modulation noise figure of a klystron, whereas HOBSON[24] sees some correlation between the frequency modulation noise and current noise. Improvement in frequency modulation noise might therefore be possible by using a constant voltage and current bias supply.

4.5 CAVITY DESIGN

Apart from frequency variations due to frequency modulation noise and temperature variations, a major contributor to the stability of the output microwave frequency is the stability of the resonant cavity itself. Various authors BOTT[7], BYBOKAS[9], CLUNIE[12], COURT[14], DAY[15], HOLLIDAY[27], ITO[29], JOCHEN[31], KOOI[33] and WARNER[48] discuss construction of suitable resonant cavities. All suggest that a waveguide cavity will give a higher Q than a co-axial cavity as well as reducing the effects of frequency modulation noise, bias supply variations, temperature variations and load admittance variations on the oscillation frequency. A waveguide cavity approximately $n\lambda g/2$ in length is favoured by CLUNIE and COURT, optimum frequency stability with suitable power output being obtained for n = 4. BYBOKAS, DAY and ITO describe the use of a reduced height waveguide as the resonant cavity. BYBOKAS in particular suggests decoupling (2 dB) to improve frequency modulation noise performance and temperature stability. With this amount of decoupling the output power and frequency will also be less affected by external load changes. Frequency pulling is also reduced (by 50%) by the decoupling and can be much further reduced by the use of an isolator between the cavity and the antenna. DAY gives a good description of the major features of cavity design, mentioning in particular that the bias line should be designed as a filter to reject the microwave energy.

With regard to internal, external and ground clutter microphony, information present in the above references indicates that the solid-state diodes together with a mechanically rigid cavity with good matching to and isolation from the antenna compare favourably with klystrons for the same power output. Minimising the effect of strong local reflections is imperative in order to avoid frequency pulling, especially for a two-oscillator CW Doppler radar.

4.6 SUMMARY

Solid-state microwave oscillators, especially Gunn diodes, would seem to have the desired characteristics for the production of a two-frequency spectrum using two separate microwave oscillators. Their stability of microwave frequency and power output is well within the limits required for a CW Doppler radar.

CHAPTER FIVE

FREQUENCY DISCRIMINATORS

5.1 INTRODUCTION

It would appear that there are three principal methods for producing a voltage proportional to the frequency difference (Δf Hz) between two separate microwave oscillators, namely 1) cavities tuned to microwave frequencies spaced Δf Hz or more, up to approximately f/Q Hz, apart, 2) RLC tuning to the difference frequency, and 3) a digital counter of difference frequency cycles together with a digital-to-analogue converter. The difference frequency of interest to the present discussion will be of the order of 1 MHz, hence methods 2) and 3) would both require a "video-frequency" amplifier. Application of the voltage produced by the discriminator will be governed by the usual principles of control theory such as negative feedback with a stable phase-versus-frequency characteristic; production of an error signal and application of a control voltage via an integrator.

5.2 CAVITIES TUNED TO DIFFERENT MICROWAVE FREQUENCIES

As mentioned in 2.5, various authors have used this method for automatic frequency control. None of them used two oscillators to transmit two microwave frequencies and the value of Δf was usually of the order of 50 MHz.

5.3 LUMPED CIRCUIT TUNED TO THE DIFFERENCE FREQUENCY

An RLC circuit can easily be tuned to a difference frequency of 1 MHz. Crystal tuning might be required to obtain the required stability with temperature variations. Incorporating the tuned circuits in the video or IF-frequency amplifier would reduce the noise in the control circuit, especially noise signals due to strong local reflections at the order of the Doppler frequency. The difference frequency would have to be amplitude limited in order to avoid amplitude variations in the output of the frequency discriminator due to amplitude variations in the difference frequency caused by Doppler signals. An FM radio receiver design could also be used to produce a voltage output proportional to the difference frequency. The continuous variation of the voltage output for the lumped circuit discriminator should cover the input ranges of Δf equal to 1 MHz $\pm$ 0.25 MHz. (Linearity is not essential). Most of the components for this type of discriminator would be integrated circuits with the exception of the inductances. HOBSON[25] gives for Δf = 750 kHz and Q = 10, a response of 20 kHz/µs which should be fast enough to follow the variations in Δf Hz that will occur for Gunn diodes.

5.4 DIGITAL FREQUENCY DISCRIMINATOR

This form of frequency discriminator was chosen for the construction of the two-oscillator radar to be described in Chapter 7, in order to study the use of integrated circuits and standard, well tried, digital methods for this type of radar. DTL counters are easily able to count at frequencies up to

2 MHz; at 10 to 100 MHz TTL and emitter-coupled logic would be required. The difference frequency signal was gated into the counter, such that the input frequency was counted for 100 µsec. 0.5 MHz would give a count of 50. Two decades were therefore required. At the end of the 100 µsec period, the input to the counter was inhibited, and by means of 2 quad-latches the count was transferred to a temporary storage register (bistable latches) in the form of 8 flip-flops. The counters were then reset, counter 1 then counter 2, and the input gate was then opened. The time to store the count and reset the counters was approximately 15 µsec, giving a sampling rate of approximately 8.5 kHz (See Fig.3 for timing information). The 8 flip-flop storage register was connected directly to a 10-bit current source (Fairchild µA 722). The resistor network was weighted 1, 2, 4, 8, 10, 20, 40, 80 for use with the decade counters (See Fig.5 for resistor values). Full scale i.e. a count of 99 would give approximately 2.5 mA output. This current was then converted to a voltage by means of the operational amplifier (µA 741) such that 0.5 MHz input to the counter gave 1.0 volts output. The slew-rate of the µA 741 (1 volt/10 µsec) is such that the frequency response of the discriminator is less than 10 kHz (See Fig.14 and Fig.15 for the discriminator output with an F.M. input). The non-linearity in the curve is due to mismatch of the resistor network and shows up most at the change over from 30 to 40 and 70 to 80. The 115 µsec sample-time can also be

seen from Fig.14. It is to be noted that the discriminator will give a voltage output for input frequencies from 10 kHz to 990 kHz.

In order to drive DTL or TTL the positive going input has to become greater than 2.5 to 3.0V to be considered high, and has to be less than 0.4V to be considered low. The difference frequency voltage signal across the crystal detector was about 10 to 100 mV, depending on the power output of the microwave diodes and the coupling between the two channels. The only integrated circuit video amplifier that could be found, at the time, that would give the required output amplitude was the RCA3001. Its specification gave 1 volt rms up to 10 MHz. Three of these amplifiers were used in series to give sufficient amplification for the complete range of difference frequency amplitudes at the crystal detector. The RC coupling between the stages gave the 3 dB point at the low frequency end of the bandwidth at 100 kHz. The amplifiers themselves gave the upper frequency limit (no other filtering was used). The RCA3001 also is emitter follower input and output which makes it suitable for driving logic integrated circuits. The differential amplifier configuration gives an AGC facility if required. $\pm$ 6V bias supply is required to give the 1V rms output (See RCA MANUAL[49]). The amplifier output was connected via a capacitor to the base of Q1, (See Fig.4), which was normally saturated. A negative going input $>$ -0.7V would switch transistor Q1 off, hence giving a high voltage on its collector. This voltage was sufficient to drive the DTL gate input. The high gain (X1000) of the 3 RCA3001 amplifiers also gave some limiting hence helped to stabilise the input to the counter when the amplitude of the difference frequency varied.

The clock frequency, approximately 10 kHz, was generated by an astable-multivibrator. Frequency adjustment was by means of a variable resistor. All other pulses were sequenced from the astable-multivibrator output by means of a delay network (See Fig.7). The amount of delay being fixed by the value of the capacitor (C2). The latch and reset pulses (Fig.3) all used the same value of capacitance (100 pF). No attempt was made to stabilise the frequency of the astable-multivibrator against temperature variations. The clock waveforms were connected to the counters and latches via DTL quad NAND-gates.

The power supply requirements were +5V at 250 mA. This was provided by a lead-accumulator; the +5V being obtained by dropping approximately 1.25V across two 10 Ω resistors in parallel. This meant, in fact, that the digital circuits had 4.75V to 5.0V depending on the state of charge of the lead-accumulator. The 10-bit current source had a separate $\pm$ 6V dry-cell bias supply.

5.5 SUMMARY

The construction of a frequency discriminator for a control system used to control the difference frequency between two microwave oscillators in the region of 1 MHz would seem to present no great technical problems. The digital method has the properties of long term stability, speed of response, high noise immunity, and ease of construction. The lumped circuit could be considered to be simpler but requires inductances which are bulky; crystals might also be required for temperature stability. The lumped circuit discriminator has a higher Q than the digital circuit and hence is less effected by other frequency signals at its input, e.g. Doppler frequency signals. No one method seems to have any great advantage over the other; the choice is the experimenter's.

CHAPTER SIX

CONTROL ELEMENT FOR SOLID-STATE MICROWAVE OSCILLATORS

6.1 INTRODUCTION

The error voltage (or current) produced by the difference between the output of the frequency discriminator and a stable reference voltage (or current) has to provide negative feedback to control the difference frequency between the two microwave solid-state oscillators. Application of this voltage (or current) will in general be via an integrator in order to give a stable difference frequency. For a klystron the control of the frequency can be performed via the repeller voltage. For solid-state microwave oscillators, on the other hand, an external element must be employed, usually to give variable loading on the waveguide cavity and hence vary the output frequency of the microwave diode plus cavity. The two most useful elements are the varactor diode and the YIG ceramic. By changing the reverse bias voltage on the varactor, its capacitance will change, and hence the effective capacitance of the microwave circuit will change together with its frequency of oscillation. A YIG ceramic gives a variable loading of the cavity, (See ELLIS[17], p53), when placed in a voltage variable magnetic field. Since variation in temperature changes the output frequency of a solid-state microwave oscillator some form of control is required to follow the slowly changing output frequency due to this variation in the temperature. Both this and bias supply variations causing output frequency variations have been mentioned in Chapter 4.2. Mechanical vibrations of the solid-state microwave diode with respect to the cavity will also cause changes in output frequency. Rigid mechanical construction is essential in overcoming this form of frequency variation with time.

6.2 VARACTOR DIODES

Variation in the reverse bias voltage applied to a varactor diode varies its capacitance. When this varactor is coupled to the cavity in the microwave circuit containing the solid-state microwave power source (See WARNER[48] for circuit details), this variation in capacitance will change the frequency of oscillation of the microwave circuit. Also the variable coupling of the varactor to the cavity causes output power variations (See HOLLIDAY[27] and WARNER[48]). COUCH[13] reports a tuning range of 1100 MHz, with a shunt connected varactor diode, the tuning range being determined by the Q_L of the varactor. WARNER gives a tuning range of 600 MHz for a $(C_o-C_v)/C_v = 0.37$ for the varactor diode. SMITH[46] obtains 3 GHz tuning for 8.6 GHz/pF, together with a linear variation of frequency with capacitance of the varactor diode; the power output variation was 1.2 mW to 1.7 mW. For a tuning range of 750 MHz in the X-band with an output power variation of 2 dB at 10 mW, ELLIS[17] mentions the point that because of the varactor loss, the circuit has a low loaded Q, and therefore might not be suitable for applications where the frequency modulation noise is important. COURT[14] points out that a high Q, $2\lambda g$ cavity will reduce the electronic tuning range. AITCHISON[3], using thin-film lumped circuits together with a Gunn diode and a varactor diode maintains that a variation in C_v of 3 to 1 should give a tuning range of two to three GHz (at 10 mW output). An improved C_v variation of the varactor was obtained by removing the encapsulation. It is to be noted, from AITCHISON, that a series connection of the varactor diode is preferred; if the varactor diode is connected in parallel, then rapid variations of output power occur as the varactor capacitance changes.

BOTT[7] and HOLLIDAY[27] note the point that the larger the coupling between the two microwave solid-state oscillators then the tighter must be the coupling between the Gunn and the varactor diodes to achieve the required frequency or phase change. For applications where speed of response is important HOLLIDAY[27] and SMITH[46] point out that the varactor diode has a much faster response than YIG ceramics.

6.3 YIG CERAMICS

FANK[18] notes the speed limitation of YIG in comparison with varactor diodes together with the fact that the variation in output microwave frequency will be linear with control voltage (or current). FANK also points out that PIN diodes could be used for discrete tuning of the output microwave frequency. BYBOKAS[9] reports an octave bandwidth tuning but describes technical difficulties in locating the YIG sphere near the Gunn diode. SMITH[46] maintains that with YIG it is possible to obtain a wide tuning range at the microwave frequency but again notes that the speed of response of YIG is much less than that of a varactor diode. ELLIS[17] makes the point that one attractive feature of YIG is its low loss (of microwave power) compared to the varactor diode, and gives the figure 2000 for unloaded Q at 10 GHz. Hence the YIG can obtain comparable tuning ranges but higher loaded Q's than the equivalent varactor diode tuning circuit. 1000 MHz tuning range is available, with power output variation from 6 to 8 mW, using strip-line circuits.

6.4 SUMMARY

The output microwave frequency of a solid-state microwave oscillator is able to be varied by a variable coupling, to the associated cavity or resonant circuit, of a varactor diode, PIN diode or YIG ceramic. The varactor diodes and YIG have comparable tuning ranges, with the YIG giving higher loaded Q's and hence less output power variation with change of microwave oscillation frequency. The varactor diode coupling, on the other hand, has a much faster frequency response than the YIG. It is to be noted that the variation in the bias voltage of a Gunn diode can be used as a control parameter to vary the output frequency of the diode, as long as the subsequent variations in the power output can be tolerated (See FANK[18], FITZSIMMONS[19] and WARNER[48]).

CHAPTER SEVEN

CONSTRUCTION AND ADJUSTMENT OF A TWO MICROWAVE OSCILLATOR, SINGLE APERTURE, DOPPLER RADAR

7.1 INTRODUCTION

The main features of a two microwave oscillator, single aperture, Doppler radar are 1) the antenna design, 2) the design of the resonant cavity, to be used with the solid-state diode, to give the required microwave frequency, 3) the microwave circuit design, including the adjustment of the amount of coupling between the oscillators, and of each oscillator to the antenna and its own mixer, 4) the design of the control circuit, including both long and short term control of the difference frequency, and 5) the Doppler channels. Of the features mentioned above number 3), the adjustment of the amount of coupling between the oscillators and of each oscillator to the antenna and its own mixer is the most difficult to achieve with consistency. Unless these couplings are correct, the proper phase relationship between the two Doppler signals, as the range of the target varies, is not observed. This would therefore, seem to be the most critical feature encountered in the design of such a radar.

7.2 ANTENNA DESIGN

The original design was for two spaced antennas, in order to isolate the two oscillators sufficiently to avoid locking. The antennas chosen were 30 cm diameter paraboloidal dishes with focal feed. This meant that the foci were spaced approximately 30 cm when the dishes were mounted one vertically above the other. Each dish was coupled via a circulator and isolator to one of the oscillators and a waveguide feed, including an attenuator, was provided between the two oscillators in order to control the amount of coupling between them. After much adjustment, two Doppler signals were obtained one from each of the respective detector crystals attached to the circulators. When displayed on an oscilloscope as Lissajous' figures, the trace obtained, (the oscillator frequencies being set about 1 MHz apart), was an ellipse which rotated five or six times as the target moved over a range of 50 metres. This multiple rotation would seem to be due to the effect of ground reflections; each antenna acts like "Young's single slit", producing its own fringes.

In order to overcome this multiple rotation of the Lissajous' figures, the next antenna configuration chosen was a single parabolic dish with the two oscillators feeding the dish via two vertical-E waveguide feeds, with their ends adjacent vertically above one another, at the focus of the paraboloid. This was considered to be the closest spacing possible with separate feeds from each oscillator. The coupling between the oscillators was now much stronger but the control circuit was able to prevent locking at a difference frequency of 0.5 MHz. Moving targets still gave a rotating Lissajous' figure. Although the rotation was less than for the 30 cm spaced antennas, it was found that the rate of rotation depended on the bearing angle of the target and on the rate of change of bearing angle; the rate of phase change still did not follow the theoretical prediction. Hence it was considered that no spacing at all could be allowed between the antennas, if one sought to obtain an estimate of the range of the target from the phase difference between the two Doppler signals. A single aperture is essential.

A possible method of feeding a single aperture with microwave power at two frequencies spaced approximately 1 MHz apart would be to have the radiation associated with one of the frequencies polarised at right angles to that of the other frequency. Two TE_{11} modes in a circular waveguide will propagate independently if orthogonally oriented. FOGEL[20] uses a cruciform assembly to couple one E-plane and one H-plane bend to a circular waveguide via a matching taper. A VSWR <1.15 with -40 dB isolation was achieved over a 10% bandwidth at X-band frequencies. The mode isolation characteristic for this type of construction is very dependent on the mechanical symmetry between the rectangular and circular waveguides. TOMPKINS[47] couples the TE_{10} mode in two separate rectangular waveguides to the TE_{11} modes in a single circular waveguide. One rectangular waveguide is coupled radially and the other is coupled, via a dielectric transformer, axially. Isolation between the two spaced junctions is provided by radial short-circuit pins and an absorbing septum. A VSWR <1.15 at frequencies 8.6 to 9.6 GHz was achieved in both rectangular arms together with a mode isolation of −50 dB. As for FOGEL, good mechanical construction is required to maintain this high order of isolation.

No attempt was made in the present work to transmit two differently polarised waves through a single aperture, although in work on an earlier version of this project some experiments were made with orthogonally-polarised separate transmitters.

7.3 RESONANT CAVITY DESIGN

Both waveguide and co-axial cavities were used with the Gunn diodes (MULLARD, 820 CXY/A), to form the microwave oscillators. The variation of frequency with temperature was observed to be less with the waveguide cavity, as was the variation of frequency with varactor coupling. The higher loaded-Q of the waveguide cavity allowed greater coupling between the oscillators before locking was observed. The co-axial cavity was finally chosen to enable ease of tuning to the assigned microwave frequency (9.25 GHz); fine tuning was obtained by moving the variable short circuits (E and F, Fig.8). The varactor diode was coupled into the co-axial cavity via a loop inside the cavity opposite the output power coupling loop (See Fig.29 and WARNER[48]). Most Gunn diodes require cooling fins to dissipate the 1 to 4 watts of input power. The bias supply had no temperature or voltage stabilisation hence the warm up period was observable both in the variation of power output and variation of current drawn by the Gunn diode as its temperature increased with time from switching on.

7.4 MICROWAVE CIRCUIT DESIGN

The block diagram of the final form of the microwave circuit is given in Fig.8 (See also Fig.30 and Fig.36). The co-axial cavities were tuned to the required frequency and power output by use of the screw-tuners G and H on the cavity itself and by the correct orientation of the output power coupling loop. The power output was monitored by measuring the voltage across a 1 kΩ resistor in parallel with the detector crystal. A typical voltage would be in the range 0.5 to 1.0 volt. The microwave power from the Gunn diode and resonant cavity was passed to the single feed antenna via the isolator, circulator, three-screw-tuner and attenuator. The isolator was included to minimise frequency pulling of the output microwave frequency due to strong local reflections. The isolator also decreases the coupling between the two Gunn diode oscillators. The circulator not only gives good Doppler frequency mixing of the transmitted and reflected microwave signals but also contributes to the isolation between the two microwave oscillators. The isolators and circulators together give an isolation of at least -40 dB in each direction between the two microwave oscillators. The attenuators A and B (Fig.8) were inserted to be able to increase this isolation if, and when, required. The tee-junction also gives -3 dB isolation in each direction between the oscillators.

The attenuation from these attenuators was set at 0 dB for all the recorded results.

The proper adjustment of the 3-screw-tuners C and D (Fig.8) was essential in order to obtain good matching of each Gunn diode oscillator to the antenna. The 3-screw-tuners, C and D, were adjusted to give maximum detector crystal current, i.e. maximum voltage across the 1 kΩ resistor in parallel with the detector crystal, while at the same time the variable short circuits, E and F (Fig.8), were also adjusted to maintain the required difference frequency i.e. the short-circuits enabled fine tuning to be achieved. The 3-screw-tuners C and D were then readjusted to give a phase difference between the two Doppler signals present, one in each detector crystal, which approached zero for local reflections from a moving target. In order to obtain zero-phase, between the above two Doppler signals, for zero-range of a moving target a microwave phase-shifter was substituted for the attenuator A. Assuming correct coupling between each oscillator and its respective mixer, introduction of a phase-shifter in the microwave circuit at the position A in the waveguide is the one position (apart from B by symmetry) which will introduce a microwave phase shift between the reflected microwave signal and the reference microwave oscillation at the detector crystal position in the waveguide (See Fig.19 for zero-phase at zero-range). The angle of the coupling loop used to couple the varactor diode to one of the

Gunn diode oscillators is to be adjusted to give the maximum microwave frequency change of that oscillator for a given reverse bias voltage change on the varactor diode. The position of this coupling loop is also to be adjusted to avoid too much microwave power absorption in the varactor diode itself. The latter adjustment therefore, limits the maximum microwave frequency change obtainable. The difference frequency was displayed continuously on an oscilloscope; the monitor point was usually the input to the counter in the frequency discriminator (See Fig.4 and Fig.10).

7.5 CONTROL CIRCUIT DESIGN

Fig.5 shows the main features of the control circuit design. The output of the current to voltage converter, monitor point (7), Fig.5, gave a linear variation from 0V to +2V such that an input difference frequency of 0.5 MHz gave an output of +1V. Since the Gunn diode was biased positively with respect to the waveguide, the tuning varactor, VAT71 (VARIAN), was reverse biased negatively with respect to the waveguide. A voltage inverter was, therefore, required such that an input difference frequency of 0.5 MHz gave a voltage of -6V at monitor point (9), Fig.5, with a variation from 0V to -12V.

A reference voltage of -6V was set by the potentiometer R1 (Fig.5) at the minus input to the integrator. The voltage from the inverter was applied to the plus input of the integrator, the error voltage, i.e. the voltage between (X) and (Y) (Fig.5), was, after amplification by the integrator, applied as the control voltage (V_{CL}) to the varactor (Monitor point (8), Fig.5). The time-constant of the integrator could be adjusted by changing the value of C (Fig.5), while the difference frequency could be adjusted $\pm$ 100 kHz by the variation in R1 available.

Two slow drifts were noticed in the value of the control voltage applied to the varactor. One of the drifts could be associated with that of the integrator itself. No input bias-current compensation circuit was provided, although the input off-set voltage null, available on the Fairchild µA 741, was used. The other drift was due to temperature variations in the Gunn diode, caused by wind air-currents flowing past the cooling fins of the diode housing. SELING[41] gives a circuit that will compensate for this type of slow temperature variation. Some form of temperature control would be required as well as varactor control of the oscillator frequency in order to keep the difference frequency constant over times shorter than and longer than the period of the highest and the lowest Doppler frequency expected. To compensate for the slow drift in the control voltage due to both these effects, the variable short-circuits, E or F (Fig.8) were adjusted slightly, from time to time, to bring the control voltage back to the -6V value (i.e. zero-error voltage at the input to the integrator). This adjustment of the short circuits was performed just before any of the photographs, (Fig.10 to Fig.28), were taken.

Another possible cause of this slow drift in the control voltage might have been due to a general drift in the frequency of the timing multivibrator in the discriminator, which was not stabilised against temperature variations (See Chapter 5.4).

7.6 DOPPLER CHANNELS

The Doppler signal amplifiers were constructed using two AC coupled audio frequency integrated circuit operational amplifiers, Fairchild μA 741, with a voltage gain of 100 to 1000, adjustable by varying the feed-back resistor on the second integrated circuit (See Fig.6). The phase difference between the two Doppler channels was approximately 12^{o} from 15 Hz to 10 kHz; variation of this phase difference with temperature was not measured. The switch S, (Fig.6), enabled one of the Doppler signal amplifiers to be used as a differential amplifier to obtain cancellation of in-phase local signals. This feature of the amplifier was not used to obtain ground clutter cancellation for a moving radar. Use of the phase-shifter to obtain zero-phase difference between the Doppler signals at zero-range was as far as the experiment with this radar went, i.e. the radar was always stationary, with moving targets. (It was not possible to move the radar because of the power cables required to run the oscilloscopes).

7.7 SUMMARY

The experience of constructing and adjusting a two oscillator Doppler radar would seem to indicate the following to be its major features, 1) a single aperture, i.e. a single feed parabolic dish antenna, is essential, 2) a Gunn diode together with either a waveguide or co-axial cavity has sufficient frequency stability for variations in temperature to be controlled to have a frequency differing by approximately 0.5 to 1.5 MHz from another similar microwave oscillator, 3) the critical adjustment is the impedance matching of the two oscillators to the single transmitting aperture and to the respective mixers, in order to give sufficient transmitted microwave power and, at the same time, to give a zero-phase difference between the two Doppler signals for local moving targets, 4) the "uncritical" adjustment of the control circuit; the long term stability ($>$1 minute) being of more relevance than the short term stability; the Gunn diode oscillator itself, has sufficient short term stability, 5) the Doppler signal amplifiers give sufficient gain for oscilloscope display of their output voltages for moving targets up to a range of 250 feet, for a Δf of 1.5 MHz.

CHAPTER EIGHT

SOME EXPERIMENTAL RESULTS OBTAINED WITH A TWO OSCILLATOR, SINGLE APERTURE, CW DOPPLER RADAR

8.1 INTRODUCTION

The experimental studies undertaken using a two oscillator single aperture CW Doppler radar to measure the range or imminence time of a moving target can be listed, thus: Did the relative phase of the two Doppler signals change with the range of the imminent target? Was this relative phase between the Doppler signals changed for receding and approaching targets? Did the phase difference between the Doppler signals approach zero when the range of the target approached zero? Did the phase difference between the Doppler signals vary linearly with the range of the target? Was the difference-frequency output sinusoidal or non-sinusoidal? What was the effect of temperature, mechanical vibrations and strong local reflections on the difference frequency? What was the maximum range of the radar? For the following results, 1) the microwave and control circuits, together with the Doppler frequency amplifiers, as shown in Fig.29, were positioned in the back of a stationary van (See Fig.36), 2) the Gunn diodes were MULLARD, 820 CXY/A, 3) typical values of Δf were 502 kHz and 1510 kHz, and 4) the moving target was a hand-held corner reflector (approximately 2.5 feet in diameter). The van was parked at the kerb, so that the power for the oscilloscopes could be obtained from a nearby building.

The Gunn diodes were specified to give 50 mW of microwave power output, for 9V bias voltage at 400 mA bias current. The difference frequency was periodically checked with a digital frequency meter; the usual measure of its value was obtained using the calibrated time-base of the monitoring oscilloscope. The corner reflector was chosen in preference to a moving car because of 1) its greater radar cross-section and 2) the ease with which photographs could be taken at specific distances marked on the roadway behind the parked van.

8.2 LISSAJOUS' FIGURES

As shown in Fig.19 to Fig.27 the Lissajous' figures obtained using the two Doppler signals is a stable ellipse showing that the two Doppler frequencies were approximately equal as predicted by the theory (BOYER[8], HANSEN[22] and SKOLNIK[43,44]). Also as the range (r) changed the phase difference ($\Delta\Phi$) between the two Doppler signals changed, this being indicated by the rotation of the displayed ellipse. (The maximum range from which detectable reflections could be observed, for a Δf of 0.5 MHz was approximately 200 feet). Substitution of a microwave phase-shifter for one of the attenuators, A or B in Fig.8, made it possible to obtain zero phase difference at small ranges (See Fig.19). Over the 160 feet range, the difference frequency of 1.5 MHz gave approximately 180° phase variation between the two Doppler signals, while over the same range the difference frequency of 0.5 MHz gave a phase variation of approximately 60°

(See Fig.31 to Fig.35).

8.3 RECEDING AND APPROACHING TARGETS

The reversal of phase-difference between the two Doppler signals expected when the sign of the radial or line-of-sight velocity changes could also be observed by the Lissajous' figures. The light spot, tracing the Lissajous' figure on the oscilloscope, changed its direction of rotation for receding and approaching targets, without changing the orientation of the ellipse (i.e. indicating that the target (a corner reflector) was at the same range while receding or approaching). Digital logic designed to measure the imminence time from the time difference between the zero-crossings of the respective Doppler signals would be required to take into account this reversal in sign of the phase-difference between the two Doppler signals. A knowledge of which Doppler signal belonged to the larger microwave frequency would also be required, for a correct calculation of the imminence time (See BOYER[8]).

8.4 DIFFERENCE FREQUENCY STABILITY

The major features of the stability of the difference frequency would seem to be, 1) FM noise of the microwave source, 2) the effect of strong local reflections, 3) the effect of the control voltage, 4) the effect of mechanical vibrations and 5) the shape of the difference frequency waveform.

8.4.1 FM noise in the Gunn diode

Fig.11 and Fig.13 show that the noise due to the microwave source increases as Δf decreases. Fig.23 and Fig.27 indicate the effect of this type of noise on the S/N ratio at large r. The difference frequency waveforms for 50 mW diodes were significantly noisier than those of a 10 mW diode.

8.4.2 Strong local reflections

With the digital frequency discriminator used, for very strong local reflections, e.g. those from a close moving corner reflector, it was observed that the reflected microwave power was large enough to cause frequency pulling of the Gunn diode oscillators (NAGANO[35]). This in its turn produced a variation, at the Doppler frequency, in the output of the discriminator. Hence frequency modulation of the controlled oscillator resulted. To overcome this difficulty, good isolation is required between the antenna and the microwave oscillator. Automatic gain control would also seem to be necessary (BOYER[8]).

Also under these circumstances, it was observed that the Doppler signal was large enough at the output of the high gain video-amplifier to amplitude modulate the difference frequency. Excessive amplitude modulation caused a miscount of the difference frequency and hence the two oscillators went out of the control range. Retuning with the variable short circuit was necessary to obtain the required difference frequency again.

In order to keep the difference frequency constant for strong local reflections, good isolation and AGC will be required. A high-pass filter is also required to remove all Doppler frequencies from the control circuit. A tuned circuit discriminator would have the advantage that an extra filter would not be required. Fig.18 shows the effect of strong local reflections on the discriminator output and on the control voltage. Note the control voltage varying at the Doppler frequency.

8.4.3 Control Voltage

The application of changes in the control voltage (V_{CL}) to the varactor, at a frequency of approximately 8 kHz, gives rise to a frequency modulation of the controlled oscillator, hence also a frequency modulation of the difference frequency. This variation in the difference frequency appears in the Doppler channels as an audio-frequency signal also at approximately 8 kHz. The amplitude of this audio-frequency signal can be reduced by increasing the time-constant of the control-integrator-circuit. The increased time-constant decreases the excursion of the control voltage, which decreases the change in the oscillator frequency, hence decreasing the change in the difference frequency. The effect of the strong local reflections is increased with increase of the time-constant. A fast response time of the control circuit is required in any attempt to keep the difference frequency constant because of frequency pulling due to strong local reflections. Perhaps a notch-filter would help to remove this audio-frequency and hence improve the S/N ratio of the Doppler channels (See Fig.16, upper-trace for a typical control-voltage waveform).

8.4.4 Mechanical vibrations

From the D/A converter output waveform of Fig.16, where 0.1V represents 100 kHz it can be seen that the frequency is short-term stable to ± 25 kHz over a period of 50 msec, with minor excursions to ± 50 kHz. Fig.17 shows the same waveform with the engine of the van running. The digital to analogue converter output waveform shows little change from that without the engine running (Fig.16), except that there are now a few more excursions to ± 50 kHz. The control-voltage now shows more variation, as is to be expected, with the engine running. From Fig.11 and Fig.12 of the 0.5 MHz difference frequency with and without the engine running, no noticeable difference can be observed between the two waveforms. Fig.10 shows the same waveform at the input to the counter circuit, no noticeable microphony is observed with the engine running. It is to be noted that the van was stationary while the engine was switched on. Slow large amplitude vibrations, produced by jumping up and down on the back of the van, also did not alter the difference frequency waveform. Tapping of the waveguide with a pencil produced the usual ringing in the Doppler channels but did not disturb the difference frequency waveform. No vibration tests were performed on the radar with the van in motion, because of the power cables required to run the oscilloscopes.

8.4.5 Shape of the Difference Frequency Waveform

As indicated in Chapter 3.1, equation (16), as the coupling between two microwave oscillators increases, the mixer output, at the difference frequency, changes from a sinusoidal waveform to one which is non-sinusoidal, until locking of the two oscillators occurs. Fig.13 shows the mixer output at a difference frequency of 1.5 MHz. The almost sinusoidal waveform indicates that the locking bandwidth would be about 1 MHz. Removal of the difference frequency control under these operating conditions usually resulted in an increase in the difference frequency. Fig.11 shows the mixer output at the difference frequency of 0.5 MHz. The non-sinusoidal waveform would suggest that the difference frequency was now within the locking-bandwidth and removal of the control voltage under these circumstances meant that the two oscillators locked.

Fig.11 and Fig.13, therefore, would seem to show that a) equation (16), Chapter 3.1, describes the behaviour of two Gunn diode oscillators whose frequency difference was of the order of their locking bandwidth, (1 MHz in this case), and b) it is possible to control the frequency of two X-band microwave oscillators, such that the difference in their frequencies is approximately the same as their locking bandwidth of 1 MHz.

8.5 GRAPHS OF $\Delta\Phi$ VERSUS r

Fig.31 shows a typical variation of $\Delta\Phi$ with r for a Δf of 1.5 MHz. The gradient of 1.03 degrees/foot is in fair agreement with that of 1.1 degrees/foot as predicted by theory. Fig.32 shows another typical variation of $\Delta\Phi$ with r for a Δf of 0.5 MHz over the same range as Fig.31, i.e. 40 to 200 feet. The correct gradient for this Δf is 0.37 degree/foot. The gradient for this particular set of measured values is 0.3 degree/foot, i.e. within 20% of the actual value.

Fig.33, Fig.34 and Fig.35 show the variation in $\Delta\Phi$ versus r for the same difference frequency of 1.5 MHz, but for three different settings of the 3-screw-tuner C (Fig.8). Fig.33 and Fig.35 show the typical non-linearity appearing at large r, while Fig.34 shows a linear relationship between $\Delta\Phi$ and r, with the correct gradient of 1.12 degrees/foot. These three graphs indicate the critical dependence of the linearity of $\Delta\Phi$ versus r upon the matching conditions in the microwave circuit of the two oscillators feeding the one aperture. This correct matching was very difficult to achieve while still preserving the correct Δf and transmission of sufficient microwave power to obtain the range of 200 feet. Adjustment of the length of the waveguide between the circulator and the detector crystal was also required, in combination with the settings of the 3-screw-tuners (C and D, Fig.8), to obtain the correct phase-difference variation with range.

When the correct impedance matching of the microwave circuit was not achieved it could be observed that the phase-difference ($\Delta\Phi$) did not vary at all with a change in r. $\Delta\Phi$

remained constant in the region of 180°. Introduction and use of a microwave phase-shifter to replace the attenuator (A or B, Fig.8), under these conditions also had no effect in changing the value of $\Delta\Phi$.

Adjustment of the 3-screw-tuners (C and D, Fig.8), in itself was never sufficient to obtain zero-phase at zero-range, the microwave phase-shifter in position A or B (Fig.8) was required.

8.6 SUMMARY

The major features of the results obtained using the two oscillator single aperture CW Doppler radar constructed by the author would seem to be: 1) the relative phase of the two Doppler signals did change with the range of the imminent target, as long as the impedances in the microwave circuit were properly adjusted, 2) this phase difference did change its sign for approaching and receding targets, 3) zero phase difference at zero range was only achieved by adjustment of the length of the waveguide, the 3-screw-tuners and the microwave phase-shifter, all in the microwave circuit, 4) especially critical adjustment of the 3-screw-tuners, to achieve the proper microwave impedance adjustment is essential to obtain a linear relationship between $\Delta\Phi$ and r, 5) the difference frequency was able to be controlled to $\pm$ 5% in the presence of temperature variations and mechanical vibrations due to engine noise. AGC would be required, together with good isolation between the oscillators and the antenna, to avoid the disturbing effect of strong local reflections, 6) the difference frequency waveform showed a non-sinusoidal characteristic for values of Δf less than 1 MHz, indicating that the difference frequency was being controlled within the locking bandwidth of the two oscillators, 7) the FM noise of the Gunn diodes and the noise due to the application of the control voltage to the varactor diode seemed to limit the range of the constructed radar to 200 feet with the size of antenna available. The simultaneous attainment of the correct Δf and large microwave power output was difficult to achieve due to the locking phenomena associated with the two coupled oscillators for such a small Δf in relation to the microwave frequency f.

CHAPTER NINE

SUMMARY

9.1 DISCUSSION OF THE RESULTS OBTAINED WITH A TWO FREQUENCY SINGLE APERTURE CW DOPPLER RADAR

The main features of the graphs of $\Delta\Phi$ versus r (Fig.31 to Fig.35) are 1) non-linearity and 2) when approximately linear, the gradient is in error. Chapter 8.5 indicates that these two features were caused by the mismatching of the oscillators, both to the single antenna and to their respective mixers (M1 and M2, Fig.8). A possible description of the conditions present in the microwave circuit (Fig.8) when the results (Chapter 8) were taken, can be obtained by considering the change in the phase-difference between the two Doppler signals as the coupling between the two mixer-oscillator units varies.

Consider the audio-frequency output of mixer 1 to be, say,

$$x_1 = \cos(pt + a) + u \cos(pt + f + r) \tag{22}$$

where a and f are phase constants of the apparatus, r is a phase term proportional to range, u is a constant giving the relative amplitude due to cross-coupling, and p is the angular Doppler frequency due to either of the oscillators. If $r > 0$, then this second cosine term derives from the higher frequency RF oscillation. For the audio-frequency output of mixer 2 we will write similarly

$$x_2 = v \cos(pt + b) + \cos(pt + g + r) \tag{23}$$

where b and g are further phase constants and v again represents cross-coupling. Here a, b, f and g, or rather their differences, are determined by the details of the apparatus. Without loss of generality we may put any one of these phase constants equal to 0; we put b = 0 in what follows.

We may write

$$x_1 = (\cos a + u \cos (f + r)) \cos pt - (\sin a + u \sin (f + r)) \sin pt \tag{24}$$

and

$$x_2 = (v + \cos (g + r)) \cos pt - (\sin (g + r)) \sin pt, \tag{25}$$

or alternatively

$$x_1 = A \cos (pt + Q) \tag{26}$$

where

$$A = ((\cos a + u \cos (f + r))^2 + (\sin a + u \sin (f + r))^2)^{\frac{1}{2}} \tag{27}$$

$$Q = \tan^{-1} ((\sin a + u \sin (f + r))/(\cos a + u \cos (f + r))) \tag{28}$$

and similarly

$$x_2 = B \cos (pt + R) \tag{29}$$

where

$$R = \tan^{-1} ((\sin (g + r))/(v + \cos (g + r))) \tag{30}$$

Now it is clear that the "instantaneous phase difference" observed during any short interval in which r remains essentially constant is

$$\Delta \Phi = R - Q \tag{31}$$

and, Q and R being given by the above expressions, the variation of $\Delta\Phi$ with r (in the range $0 < r < \pi$) may be computed for specified values of a, f, g, h and k.

Note first, as a check, that if u = v = 0 (no cross-coupling)

$$Q = \tan^{-1} (\sin a/\cos a) = a \tag{32}$$

$$R = \tan^{-1} (\sin (g + r)/\cos (g + r)) = g + r \tag{33}$$

so $$\Delta\Phi = R - Q = (g - a) + r ; \tag{34}$$

whereas if, on the other hand u = v = ∞ (only <u>cross</u> coupling)

$$Q = \tan^{-1} (\sin (f + r)/\cos (f + r)) = f + r \tag{35}$$

$$R = \tan^{-1} 0 = 0 \tag{36}$$

so $$\Delta\Phi = R - Q = (-f) - r ; \tag{37}$$

and finally, if u = v = 1 (no selectivity in mixer responses)

$$Q = \tan^{-1} ((\sin a + \sin (f + r))/ (\cos a + \cos (f + r))) \tag{38}$$

$$= \tan^{-1} (2 \sin ((a + f + r)/2)/ 2 \cos ((a + f + r)/2)) \tag{39}$$

$$= (a + f + r)/2 \tag{40}$$

and $$R = (g + r)/2 \tag{41}$$

so $$\Delta\Phi = R - Q = -(a + f - g)/2, \text{ independent of } r \tag{42}$$

The no cross-coupling condition, u = v = 0, shows that the phase-difference, $\Delta\Phi$, varies linearly with change in r and is zero for r = 0. The opposite condition, u = v = ∞, i.e. when only oscillator 1 is providing the LO for mixer M2, and vice versa, shows that the phase-difference, $\Delta\Phi$, will change linearly with variation of r in the opposite sense to that for no cross-coupling. (The Lissajous' figures would rotate in the opposite direction for these two conditions). The case of $u = v = 1$, gives no phase-difference change with r, i.e. each oscillator is providing the same LO power in each mixer, hence the Doppler signal output from each mixer will be the same.

Values of $\Delta\Phi$ for various values of r, have been calculated using a digital computer. The listings obtained confirmed that the linearity of the continuous variation of the phase-difference, $\Delta\Phi$, with variation in r improved as $u = v$ approached zero (and as it approached infinity) from the value 1.0. Thus, as long as oscillator 1 predominated over oscillator 2 as the LO in the mixer M1, or vice versa, then the correct linearity between $\Delta\Phi$ and r would be obtained. Suitable values of u and v could possibly be <0.1 or >10.0 to give the required linearity.

Variations in u and v were usually made, in practice, by adjustment of the 3-screw-tuners, C and D (Fig.8). Adjustment of the 3-screw-tuner C, for example, would result, most likely, in more oscillator 1 power being reflected towards the circulator, J, and from there into mixer M1, thus increasing the ratio of oscillator 1 power to oscillator 2 power in that mixer. This type of adjustment of the 3-screw-tuners (C and D, Fig.8) would, therefore, improve the linearity of the $\Delta\Phi$ versus r graphs. This was found to be so, see Chapters 7.4 and 8.5, also Figs.33, 34 and 35.

If, when the setting of the 3-screw-tuner, C, improved the LO mixing in mixer M1 due to oscillator 1, (to the extent that $\Delta\Phi$ changed linearly with r), a microwave phase-shifter was substituted for the attenuator A (position Q, Fig.9) it was observed that with adjustment of this phase-shifter one could obtain $\Delta\Phi$ equal to zero for short range targets (see Chapter 7.4). This would seem reasonable, for a phase shift introduced into the microwave signal at position Q, (Fig.9), would only effect the transmitted and reflected waves and not the LO signal from oscillator 1 to the mixer M1, and hence would control the phase of the Doppler signal in mixer M1.

Substitution and adjustment of the phase-shifter in the positions P or R, (Fig.9), did not change the value of $\Delta\Phi$, thus indicating that the rotations of the Lissajous' figures shown in the photographs corresponded to oscillator 1 being the local oscillator for mixer M1. For cross-coupling to be predominent, a phase-shifter in position Q, (Fig.9), would effect both the LO signal from oscillator 2 and the reflected waves from the antenna equally, as far as mixer M1 was concerned. Hence any adjustment of the phase-shifter would not change the value of $\Delta\Phi$. Placing the phase-shifter in the waveguide from the T-junction to the antenna was not tried.

In order to avoid this non-linearity and incorrectness in the change of $\Delta\Phi$ with change in r, a better form of T-junction would have to be used to connect the two separate oscillator-mixer units to the single antenna; the transducers proposed by FOGEL[20] and TOMPKINS[47] could, perhaps, be tried (see Chapter 7.2).

9.2 COMMENT ON THE USE OF A THREE-FREQUENCY MICROWAVE SPECTRUM IN A CW DOPPLER RADAR

A CW Doppler radar, with a two-frequency microwave spectrum, has two applications; 1) the phase-difference between the Doppler signals can be taken as a measure of the range of the target and 2) because the phase-difference between the Doppler signals is zero at zero-range, the two Doppler signals can be the input to a differential amplifier and give cancellation of local return signals. 1), above, gives range information, while 2) gives clutter cancellation of local return signals but does not give range information (except by the amplitude of the resultant single Doppler signal). It is to be noted that in application 2), as the range of the target increases, the cancellation of the differential amplifier will decrease as the phase-difference of the Doppler signals increases towards 180°. Any further increase in range will increase the cancellation again as the phase-difference approaches 360°. Hence the maximum range, for a given Δf, is limited to phase-difference, $\Delta\Phi$, of $\pm 180^\circ$ (See also BOYER[8], for the same unambiguous range). The inability of a CW Doppler radar with a two-frequency microwave spectrum to obtain, simultaneously, range information and cancellation of strong local reflections, can be overcome using a three-frequency microwave spectrum. MILNER[39] shows that transmitted microwave power, having a three-frequency spectrum, will give three Doppler signals whose phase relationship is such that they can be input, in pairs, to two differential amplifiers to cancel strong local reflected signals while the outputs of the two differential amplifiers will have a phase-difference which is a linear function of the range r. The three-frequency spectrum could consist, for example, of frequency f, f + 0.7 MHz and f - 0.5 MHz (where f = 10 GHz); these difference frequencies would give a phase-difference between the outputs of the differential amplifiers equivalent to a single difference frequency of 0.6 MHz (i.e. the average of the 0.5 MHz and the 0.7 MHz difference frequencies).

Production of a three-frequency microwave spectrum using automatic difference frequency control of three separate oscillators feeding a single aperture would seem, judging by the experience gained using this method to generate a two-frequency spectrum, to be a very complex and rather impractical method for application to a mobile CW Doppler radar. A more reasonable approach might be to frequency modulate a SINGLE microwave oscillator with a three step staircase waveform. The oscillator could be a reflex klystron, which can be frequency modulated using the repeller terminal or a Gunn or Impatt diode which can be frequency modulated by a variable loading of the cavity using a varactor diode or modulation of the bias voltage. (Care should be taken to avoid amplitude modulation when frequency modulating the microwave oscillator). The frequency of the staircase waveform would have to be greater than the largest Doppler frequency expected. Instead of tuning to the three difference frequencies as in MILNER[39], a synchronous switch would be used, as in BOYER[8], such that the Doppler signals relevant to each microwave frequency would always appear on their respective channels, with the switching frequency filtered out (see Fig.37). These three Doppler signals could then be processed, as in MILNER[39], to give simultaneous local return cancellation and range information (see Fig.37, units (13),(14) and (15)). A SINGLE oscillator would easily avoid all the locking and matching difficulties encountered using automatic difference frequency control of two microwave oscillators as described in Chapters 7 and 8, while recent developments in integrated circuits would seem to be able to give amplifiers whose phase and gain variation, with frequency and temperature changes, remain essentially the same for the expected Doppler frequencies.

9.3 CONCLUSION

Further work on a short range CW Doppler radar should, therefore, progress along the lines mentioned in 9.2 above. Recent developments in integrated circuit technology, at microwave, video and audio frequencies, suggest that it would not be too difficult to produce a three-frequency microwave spectrum by frequency modulating ONE oscillator, feeding ONE antenna, and hence make use of this spectrum's property of simultaneous clutter cancellation and range measurement. Calculation of the imminence time for a single target would then be achieved in the case when the radar itself was also moving, i.e. either as 1) a rotating antenna only, or 2) a fixed antenna on a moving platform, or 3) a rotating antenna on a moving platform.

REFERENCES

1. ADLER R.

A Study of Locking Phenomena in Oscillators.

Proc. I.R.E. and Waves and Electronics, June 1946.

2. AIGRAIN P.R. and WILLIAMS E.M.

Pseudosynchronisation of Amplitude Stabilised Oscillators.

Proc. I.R.E. and Waves and Electronics, June 1948.

3. AITCHISON C.S. and NEWTON B.H.

Varactor-Tuned X-band Gunn Oscillator using Lumped Thin Film Circuits.

Electronics Letters, Vol.7, No.4, p93, 25th February 1971.

4. ALBRECHT P. and BECHTELER M.

Noise Figure and Conversion Loss of Self-Excited Gunn diode Mixers.

Electronics Letters, Vol.6, No.11, 28th May 1970.

5. BACK I.

F.M. Noise in an Injection-Locked Oscillator when Reverse Locking exists.

Electronics Letters, Vol.7, No.12, p346, 17th June 1971.

6. BIRD J., et. al.

Gunn diodes with Improved Frequency - Stability/ Temperature Variations.

Electronics Letters, Vol.7, No.11, 3rd June 1971.

7. BOTT I.B. and HOLLIDAY H.R.

Effects of Changes in Operating Conditions on the Phase of a Frequency-Locked Gunn Oscillator in X-band.

Electronics Letters, Vol.6, No.7, 2nd April 1970.

8. BOYER W.D.

A Diplex Doppler Phase Comparison Radar.

I.E.E.E. Trans. on Aerospace and Navigational Electronics, March 1963.

9. BYBOKAS J. and FARRELL

The Gunn Flange, a Building Block for Low Cost Microwave Oscillators.

Electronics, 1st March 1971.

10. CAWSEY D.

Varactor-Tuned Gunn Effect Oscillators.

Electronics Letters, Vol.6, No.8, 16th April 1970.

11. CLEGG J.E. and CROMPTON J.W.

Low-Power C.W. Doppler Navigation Equipment.

I.E.E. Part B, Suppl.9, p258, March 1958.

12. CLUNIE D.M. et. al.

F.M. Noise Measurements on Silicon Impatt Oscillators.

Electronics Letters, Vol.7, No.2, 28th Jan. 1971.

13. COUCH L.W.

A Study of a Driven Oscillator with F.M. Feed-back by use of a Phase-Lock-Loop Model.

I.E.E.E. Trans. on Microwave Theory and Techniques, Vol. MTT-19, No.4, April 1971.

14. COURT W.P.N. et. al.

Reduction of F.M. Noise from Gunn Oscillators.

Electronics Letters, Vol.3, No.12, Dec. 1967.

15. DAY W.R.

Stabilisation of Microwave Oscillators by Injection Phase-Locking.

The Microwave Journal, Vol.10, No.4, p35, March 1967.

16. DeSa B.A.E. and HOBSON G.S.

Thermal Effects in the Bias Circuit Frequency Modulation of Gunn Oscillators.

I.E.E.E. Trans. on Electron Devices, Vol. ED-18, No.8, p557, August 1971.

17. ELLIS D.J. and GUNN N.W.

Stripline Gunn Oscillators are Compatible with Microwave Integrated Circuits.

Electronic Engineering Vol.43, No.519, May 1971.

18. FANK B.

Bulk-Effect Oscillators give Low-Cost Microwaves.

Microwaves, Vol.10, No.2, February 1971.

19. FITZSIMMONS G.W.

Microwave Generation with Avalanche Diodes and Two-Valley Gunn and LSA Mode Devices.

The Microwave Journal, Vol.11, No.1, Jan. 1968.

20. FOGEL R.L.

An Orthogonal Mode Transducer.

Nat. Conv. Rec. I.R.E., Pt.5, p53, 1956.

21. GAYLORD T.K. et. al.

Gunn Effect Bibliography Supplement.

I.E.E.E. Trans. on Electron Devices, Vol. ED-16, No.5, May 1969.

22. HANSEN W.W.

C.W. Radar Systems, Radar System Engineering, Editor Ridenour L.N., M.I.T. Radiation Lab. Series, Chap.5, McGraw Hill, 1947.

23. HARVEY A.F.

Measurement and Stabilisation of Frequency.

Microwave Engineering, Chap.19.2, Academic Press, 1963 (London).

24. HOBSON G.S.

Current Fluctuations Caused by Frequency Variations in Gunn diodes.

Electronics Letters, Vol.6, No.5, 5th March 1970.

25. HOBSON G.S.

Measurement of the Variation of Frequency with the Ambient Temperature of Microwave Semiconductor Oscillators.

Journal of Physics E, Scientific Instruments, Vol.3, p801, 1970.

26. HOBSON G.S. and THOMAS M.

Direct Frequency Demodulation with Frequency-Locked Gunn Oscillators.

Electronics Letters, Vol.7, No.3, 11th Feb. 1971.

27. HOLLIDAY H.R.

The Effect of Operating Parameters and Oscillator Characteristics upon the Phase Angle between a Locked X-band Gunn Oscillator and its Locking Signal.

I.E.E.E. Trans. on Electron Devices, Vol. ED-17, No.7, July 1970.

28. HUNTOON R.D. and WEISS A.

Synchronisation of Oscillators.

Proc. I.R.E., December 1947.

29. ITO Y. et. al.

Cavity Stabilised X-band Gunn Oscillator.

I.E.E.E. Trans. on Microwave Theory and Techniques, Vol. MTT-18, No.11, November 1970.

30. JETHWA C.P. and GUNSHOR R.L.

Circuit Characterisation of Waveguide Mounted Gunn Effect Oscillators.

Electronics Letters, Vol.7, No.15, 29th July 1971.

31. JOCHEN P.

Injection Phase Locking of Impatt and Gunn Oscillators.

Microwave Journal, Vol.14, No.2, February 1971.

32. KINGDON B.E.

A Circular-Waveguide Magic Tee and its Application to Highpower Microwave Transmission.

J. Brit. I.R.E., 13, p275, 1953.

33. KOOI P.S. and WALSH D.

Novel Technique for improving the Frequency Stability of Gunn Oscillators.

Electronics Letters, Vol.6, No.4, 19th Feb. 1970.

34. OWSTON C.N.

Automatic Frequency Control for an Electron Spin Resonance Spectrometer.

J. Sci. Instrum, Vol.41, p698, 1964.

35. NAGANO S. and AKAIWA Y.

Behaviour of a Gunn diode Oscillator with a Moving Reflector as a Self-Excited Mixer and a Load-Variation Detector.

I.E.E.E. Trans. on Microwave Theory and Techniques Vol. MTT-19, No.12, p906, 1971.

36. MARTIN B. and HOBSON G.S.

High Speed Phase and Amplitude Modulation of Gunn Oscillators.

Electronics Letters, Vol.6, No.8, 16th April 1970.

37. MATSUNO K.

F.M. Noise in a Gunn Effect Oscillator.

I.E.E.E. Trans. on Electron Devices, Vol. ED-16, No.12, December 1969.

38. MILNER C.J.

Collision Warning Devices for Road Vehicles.

I.R.E. (Australia), Radio and Electronic Engineering Conv., Melbourne, Paper No.8, May 1963.

39. MILNER C.J. (UNISEARCH LTD.)

Memorandum for Provisional Protection, Improvements in and relating to Doppler Radar (1969), Australian Patent Application, No. 0134/70, 21st January 1970.

40. SCHUNEMANN K. and SCHIEK B.

Influence of a Transmission Line on the Noise Spectra of Cavity-Stabilised Oscillators.

Electronics Letters, Vol.7, No.22, p659, 4th November 1971.

41. SELING T.V.

Thermistor Stabilises Gunn Oscillator.

Electronics, 26th October 1970.

42. SJOLUND A.

Noise in Impatt Oscillators at Large R.F. Amplitudes.

Electronics Letters, Vol.7, No.7, 8th April 1971.

43. SKOLNIK M.I.

C.W. and F.M. Radar, Radar Handbook, Chapter 16, McGraw Hill, 1970.

44. SKOLNIK M.I.

C.W. and F.M. Radar, Introduction to Radar Systems, Chapter 3, McGraw Hill, 1962.

45. SMITH M.J.A.

Microwave Discriminator for the Frequency Stabilisation of a Reflex Klystron.

J. Sci. Instrum, Vol.39, p127, 1962.

46. SMITH R.B.

Varactor-Tuned Gunn Effect Oscillator.

Electronics Letters, Vol.6, No.5,
5th March 1970.

47. TOMPKINS R.D.

A Broadband Dual-Mode Circular Waveguide Transducer.

Trans I.R.E., MTT-4, p181, 1956.

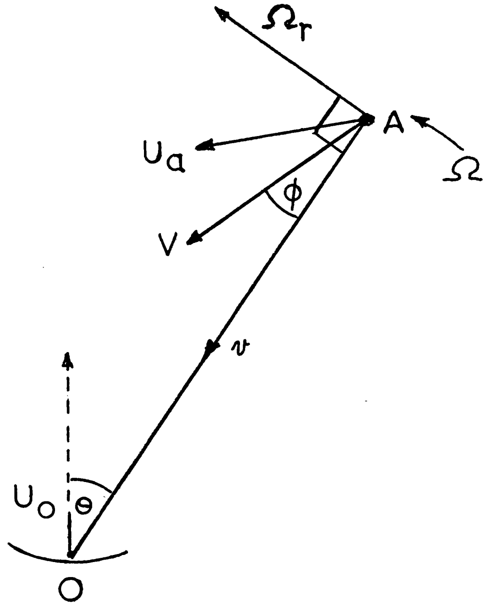
48. WARNER F.L. and HERMAN P.

Miniature X-band Gunn Oscillators.

R.R.E. Memorandum, No.2394,
H.M. Stationery Office, 1967.

49. R.C.A. LINEAR INTEGRATED CIRCUITS

Technical Series IC-41, page 145, 1967.



O = Radar Antenna A = Target

OA = range or radius r (m)

v = radial or line of sight velocity (m/s)

V = relative velocity (m/s)

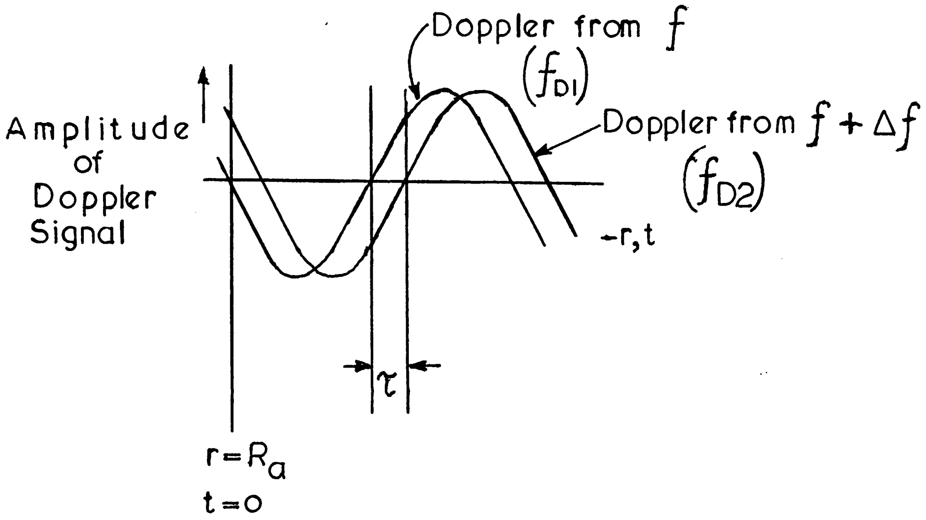
U_o, U_a = velocities of O and A, relative to some fixed point

Ω = angular velocity of A relative to O (radians/s)

Ω_r = transverse relative velocity (m/s)

(From MILNER[38], Fig.2)

Fig. 1. Various Radar Parameters

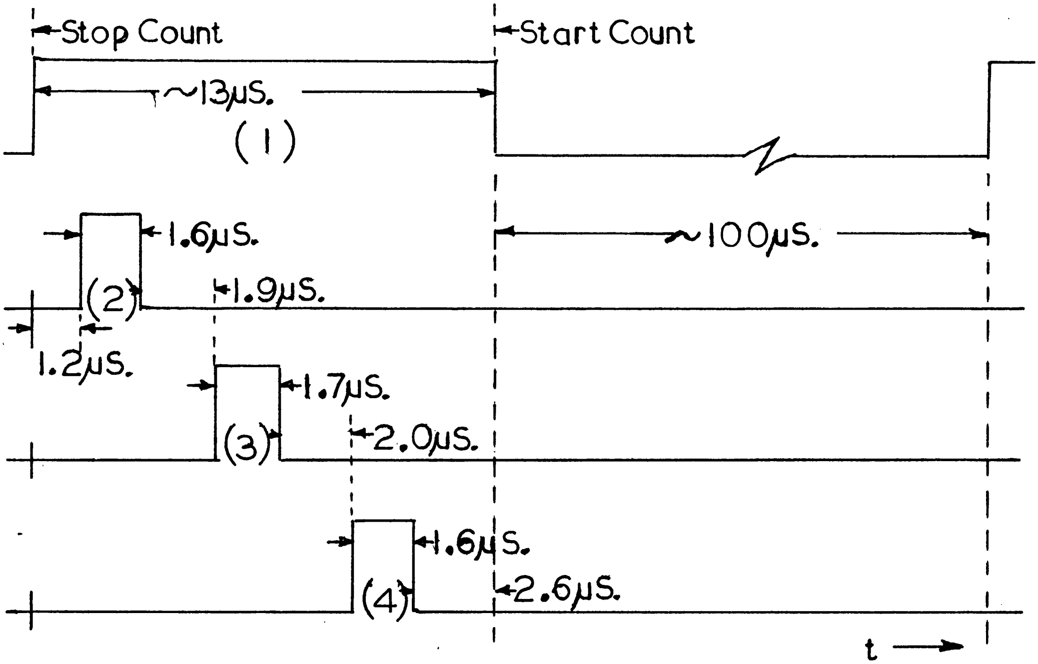


Target approaches from R_a at $t = 0$

τ = time difference between the respective zero-crossings of the two Doppler signals with frequencies f_{D1} and f_{D2}

(From BOYER[8], Fig.2(c))

Fig. 2. Phase Difference between the two Doppler Signals associated with the microwave frequencies f and $f + \Delta f$.



(1) Input Gate Waveform

(2) Store Count on Bistable Latches

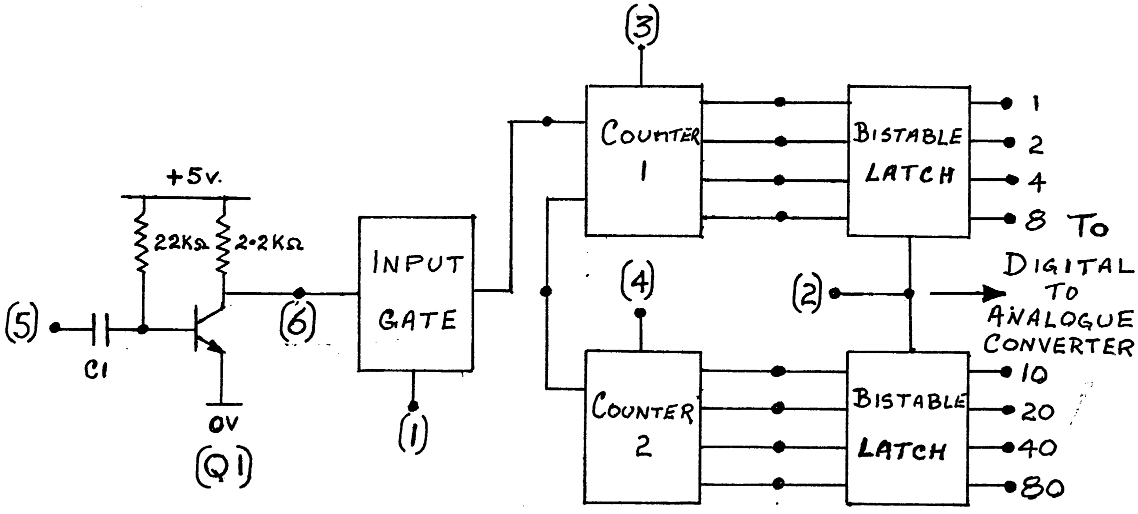
(3) Reset Counter 1, Digits 0 to 9

(4) Reset Counter 2, Digits 10 to 90

Pulse Amplitudes : +5.0 Volts

Pulse Rise and Fall Times : as for DTL

Fig. 3. Frequency Discriminator, Clock Waveforms



(1), (2), (3), (4) ≡ Pulse waveforms as in Fig. 3.

Input Gate, Decade Counters ≡ DTL (9946, 9958 Fairchild).

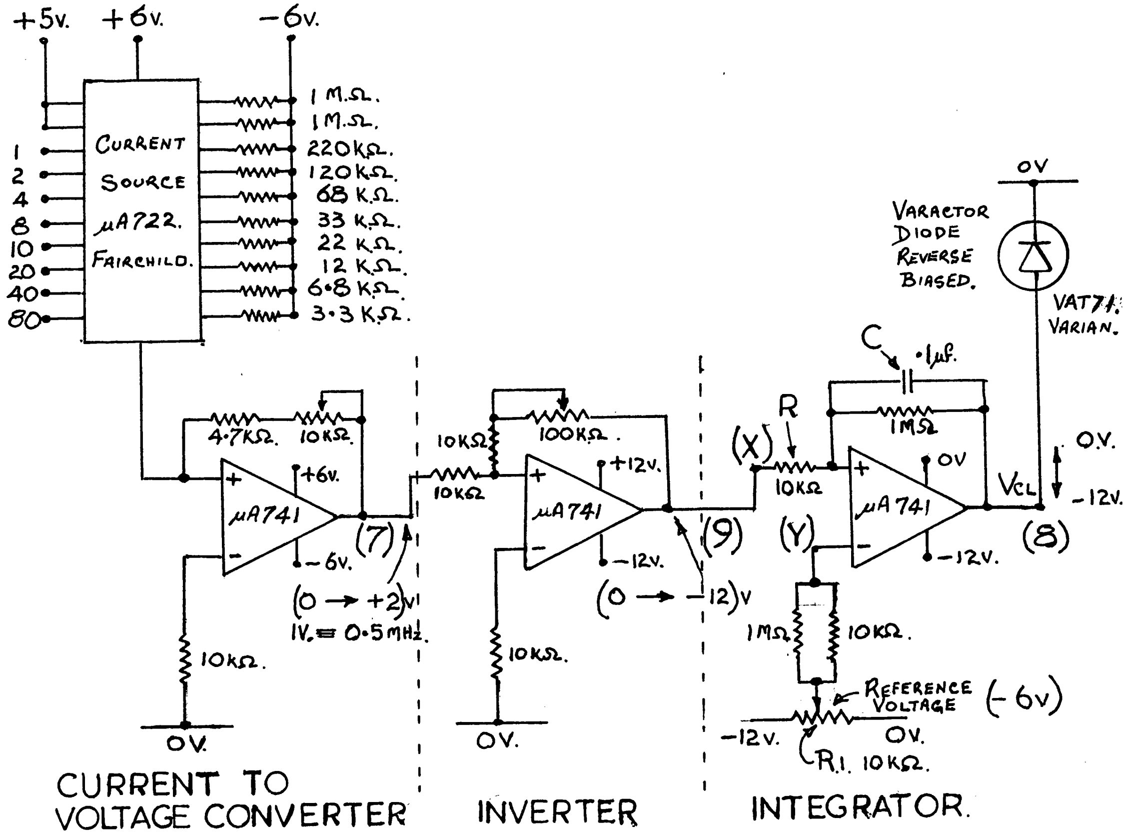
Bistable Latch ≡ TTL (7475 Texas Instruments).

(5) ≡ Video Amplifier Output (1V r.m.s. at 0.5 and 1.5 MHz).

(6) ≡ Input to the Counter Circuit (See Fig.10).

(Q1) ≡ 2N3646 (Fairchild)

Fig. 4. Frequency Discriminator, Counter/Latch Circuit



(X) - (Y) = Error Voltage

RC = Integrator Time Constant

Δf set by R1

Integrator Voltage Gain (zero-frequency) = 100

Fig. 5. Frequency Discriminator, Digital to Analogue Convertor and Control Circuit Integrator.

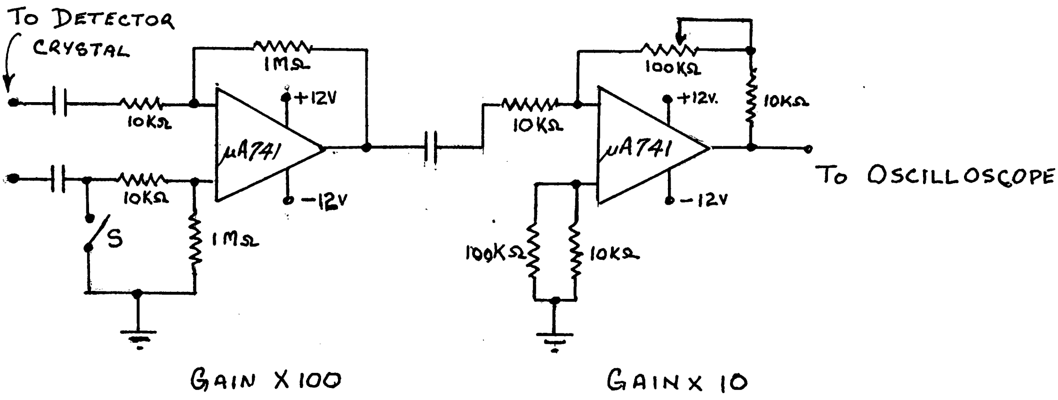


Fig. 6. Doppler Signal Amplifier

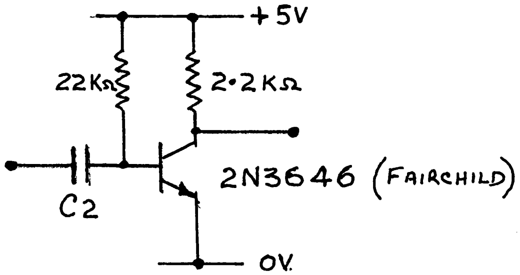
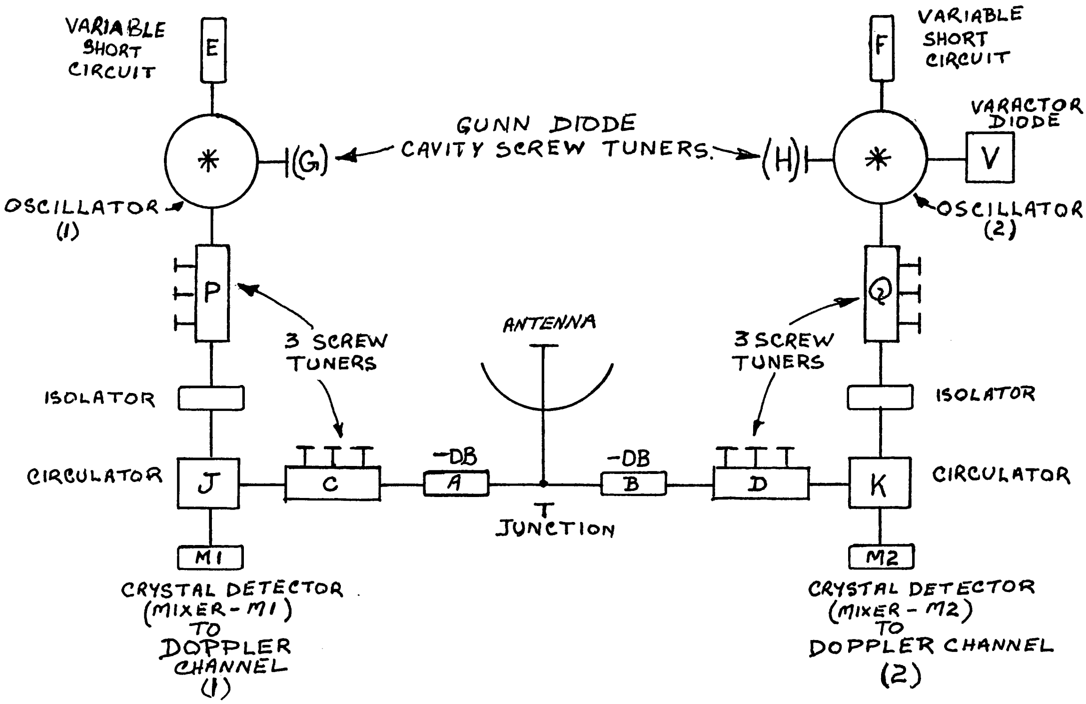


Fig. 7. Delay Network



Note: 1) All the above components are interconnected via X-band waveguide (See Figs. 29, 30 and 36).

2) ✳ ≡ MULLARD GUNN DIODES (820 CXY/A) in resonant co-axial cavity.

3) V ≡ VARACTOR DIODE (VARIAN - VAT 71)

4) The 3-screw-tuners, P and Q, were usually set with the screws fully withdrawn.

Fig. 8. Microwave Circuit Block Diagram

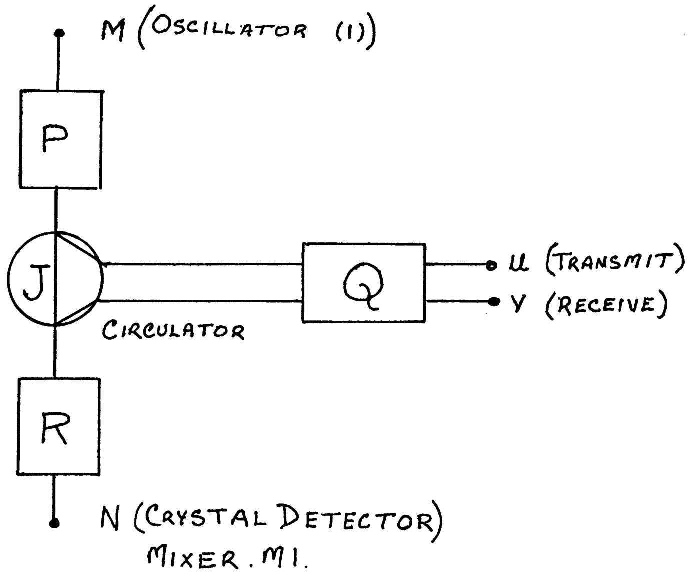


Fig. 9. Microwave Phase-Shifter Position

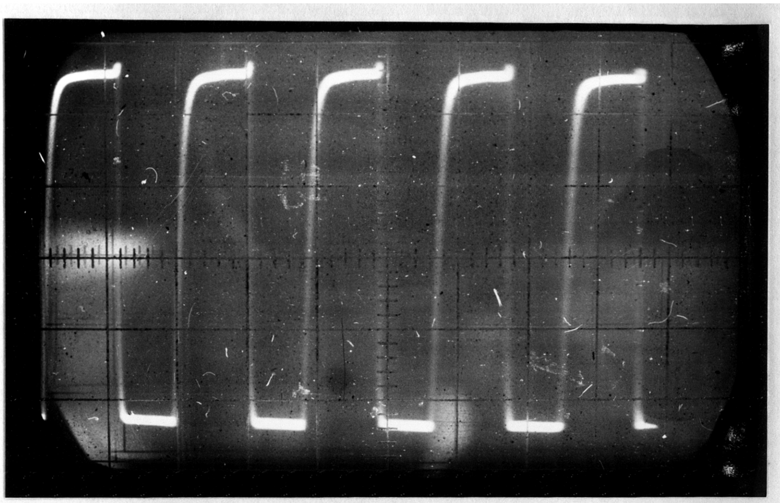


Fig. 10. $\triangle$ f input to the counter circuit, with the engine running. (0.5 MHz, 1.0V/cm, 1 μsec/cm).

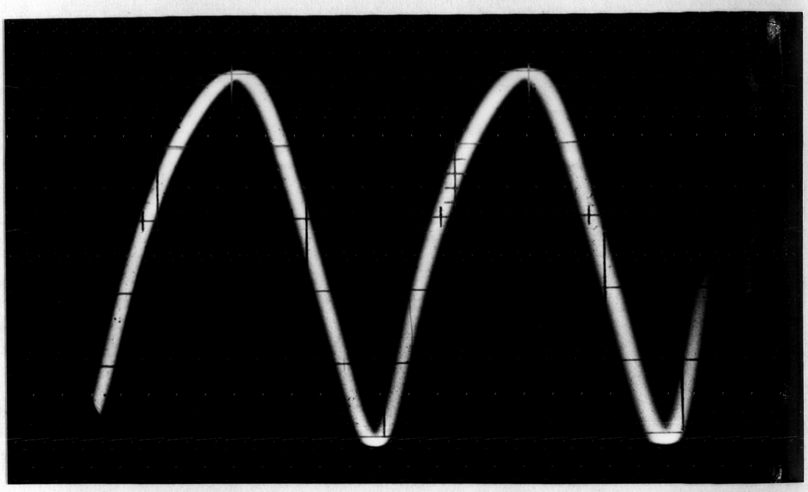


Fig. 11. $\triangle$ f input to the video amplifier, without the engine running. (0.5 MHz, 10 mV/cm, 0.5 μsec/cm).

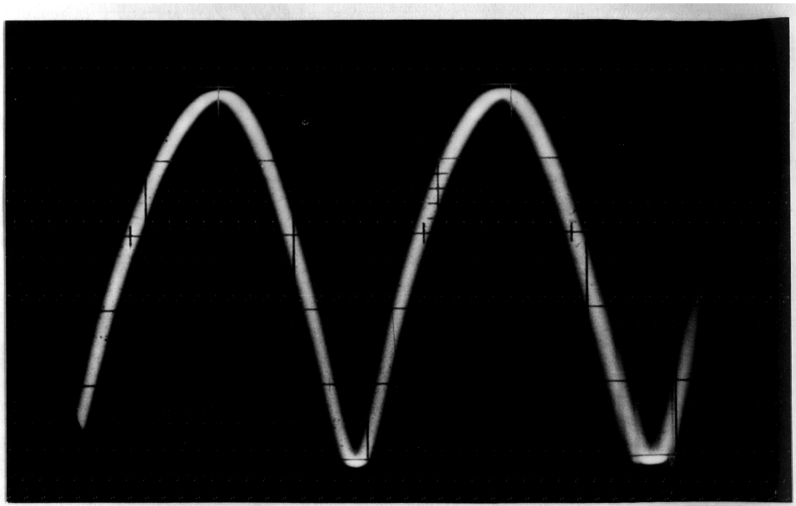


Fig. 12. $\triangle$f input to the video amplifier, with the engine running. (0.5 MHz, 10 mV/cm, 0.5 μsec/cm).

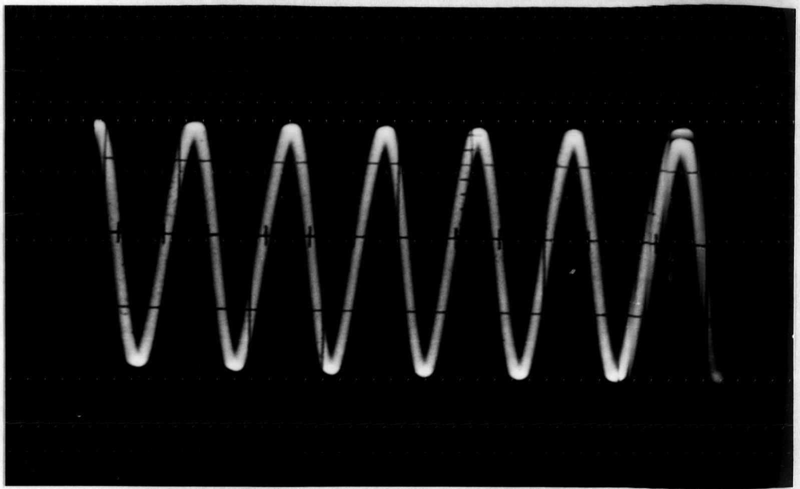


Fig. 13. $\triangle$f input to the video amplifier, without the engine running. (1.5 MHz, 10 mV/cm, 0.5 μsec/cm).

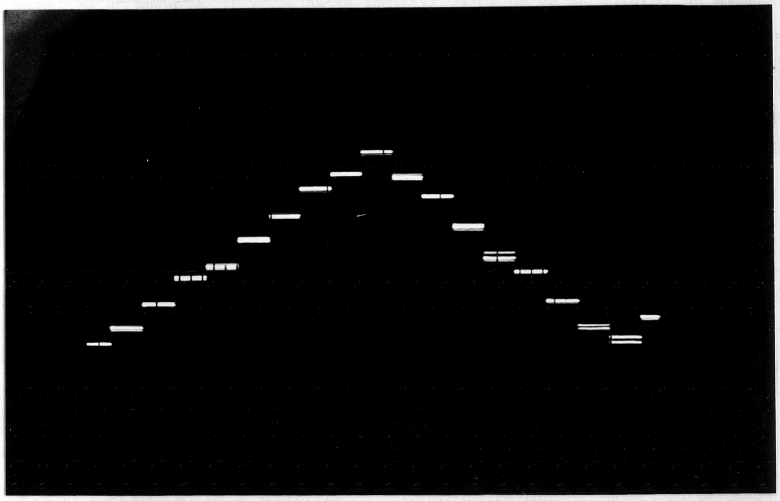


Fig. 14. Frequency discriminator output at monitor point (7) Fig.5, for FM input at monitor point (5) Fig.4. (0.5V/cm, 0.2 msec/cm).

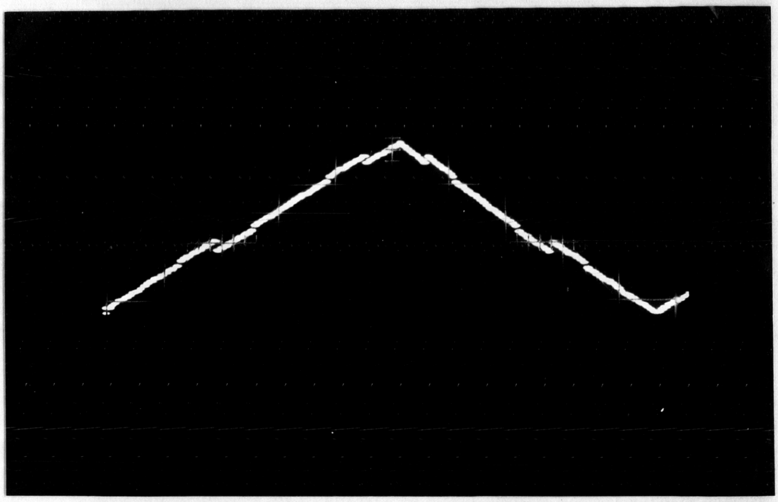


Fig. 15. Frequency discriminator output at monitor point (7) Fig.5, for FM input at monitor point (5) Fig.4. (0.5V/cm, 2 msec/cm).

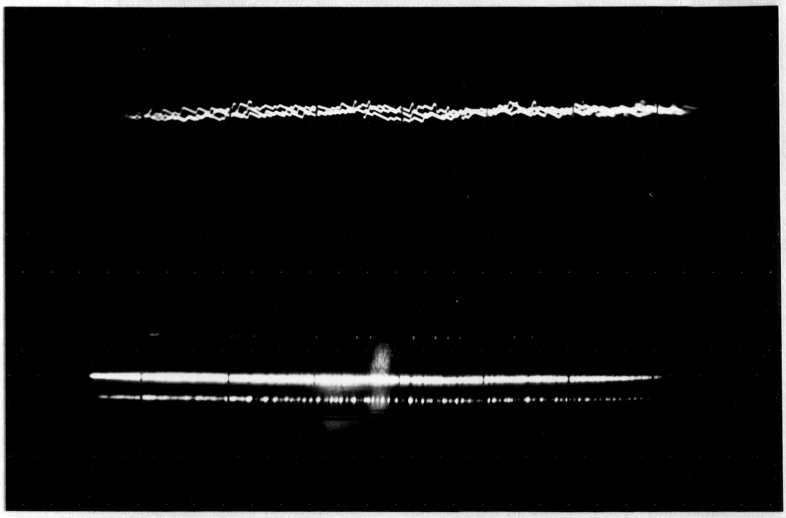


Fig. 16. Upper Trace: Control voltage at monitor point (8) Fig.5 (0.5V/cm). Lower Trace: Digital to analogue converter output at monitor point (7), Fig.5 (0.1V/cm). Δf = 0.5 MHz, without the engine running.

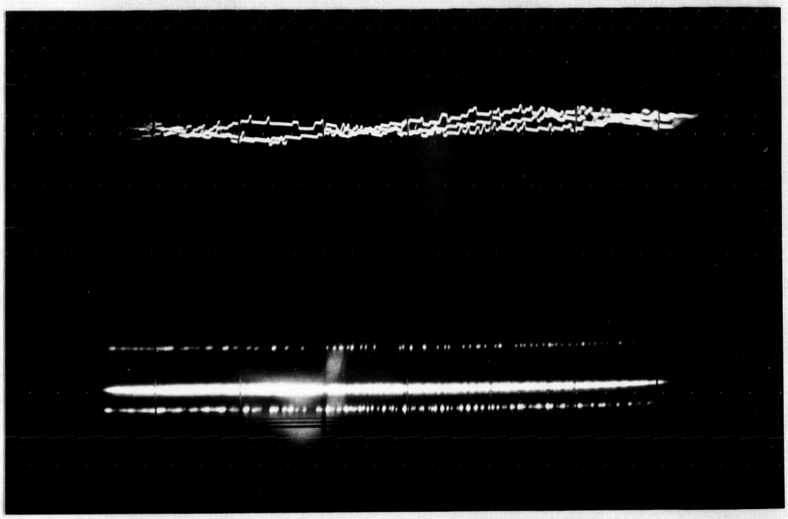


Fig. 17. Experimental conditions as for Fig.16, except with engine running.

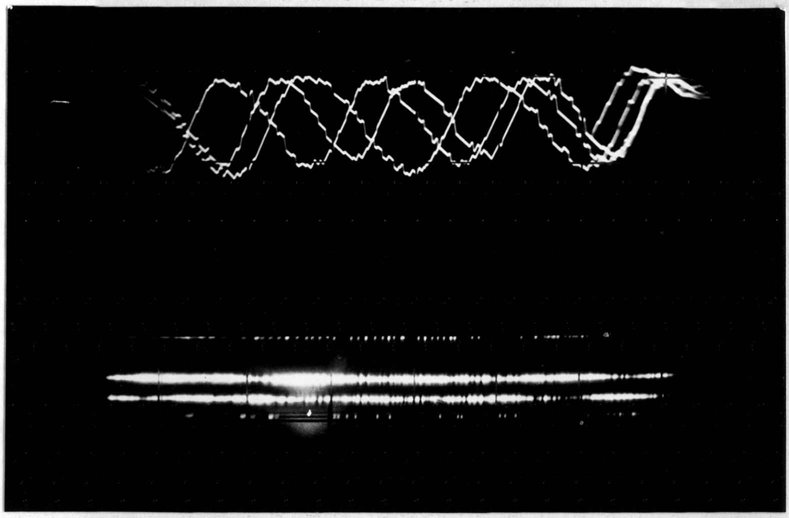


Fig. 18. Experimental conditions as for Fig.16, except with strong local reflections present.

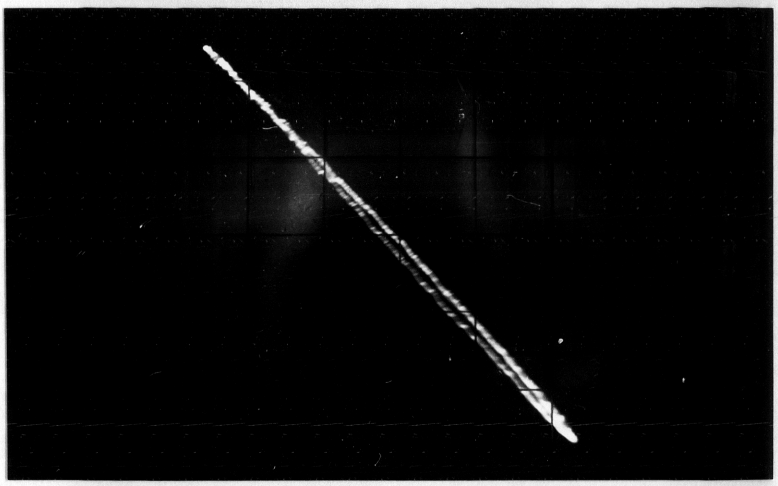


Fig. 19. Lissajous' figure, Δf = 1.5 MHz, r = 0 feet.

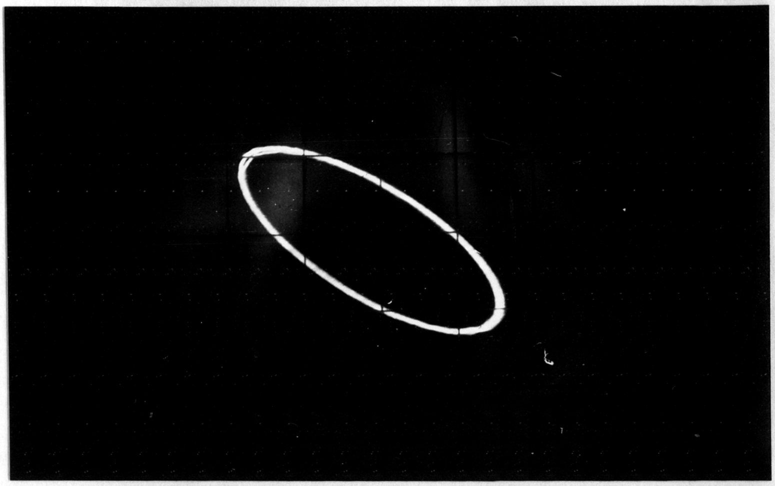


Fig. 20. Lissajous' figure, Δf = 1.5 MHz, r = 40 feet.

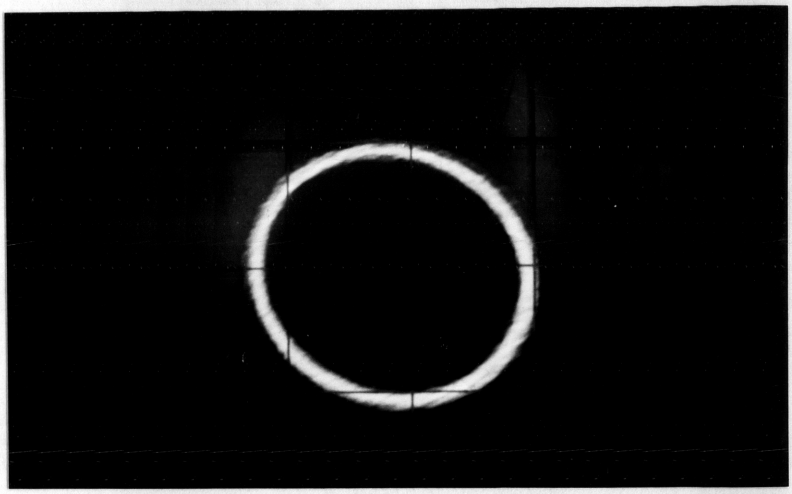


Fig. 21. Lissajous' figure, Δf = 1.5 MHz, r = 80 feet.

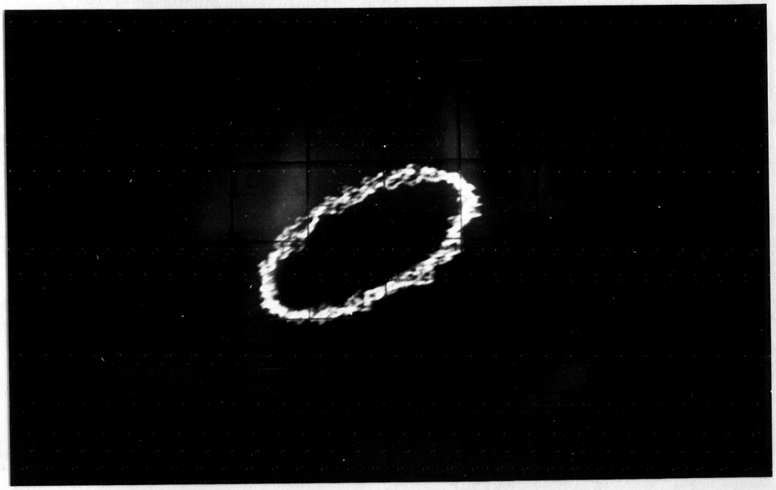


Fig. 22. Lissajous' figure, Δf = 1.5 MHz, r = 120 feet.

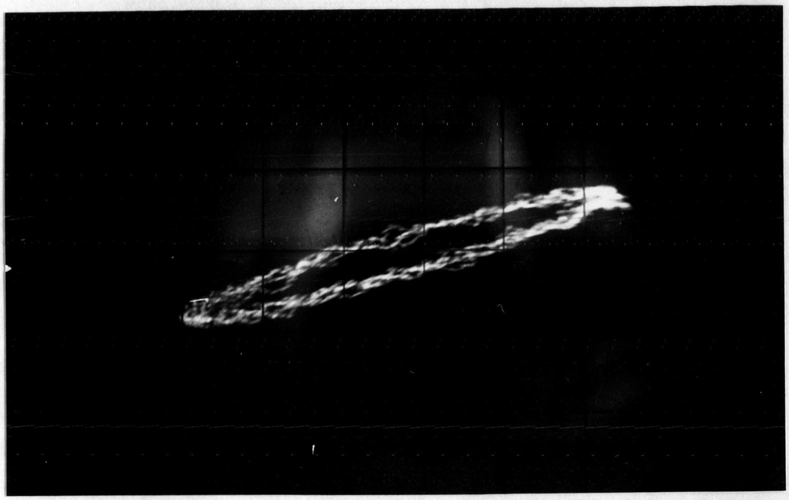


Fig. 23. Lissajous' figure, Δf = 1.5 MHz, r = 160 feet.

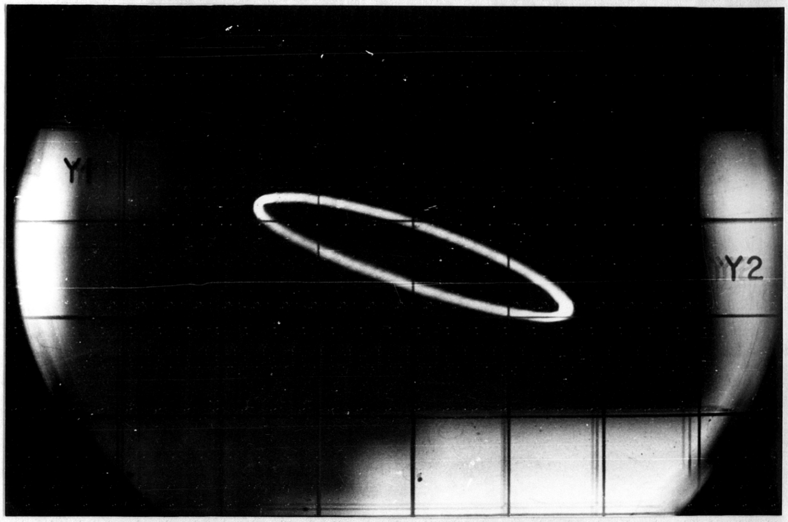


Fig. 24. Lissajous' figure, Δf = 0.5 MHz, r = 40 feet.

Fig. 25. Lissajous' figure, Δf = 0.5 MHz, r = 80 feet.

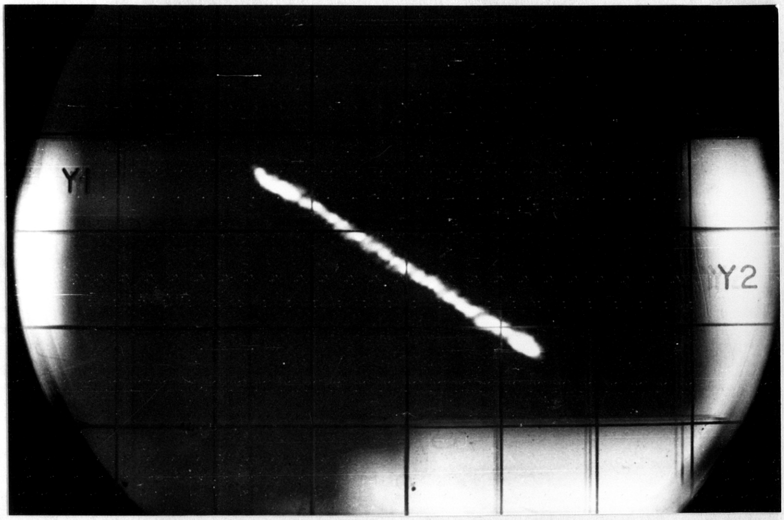


Fig. 26. Lissajous' figure, Δf = 0.5 MHz, r = 120 feet.

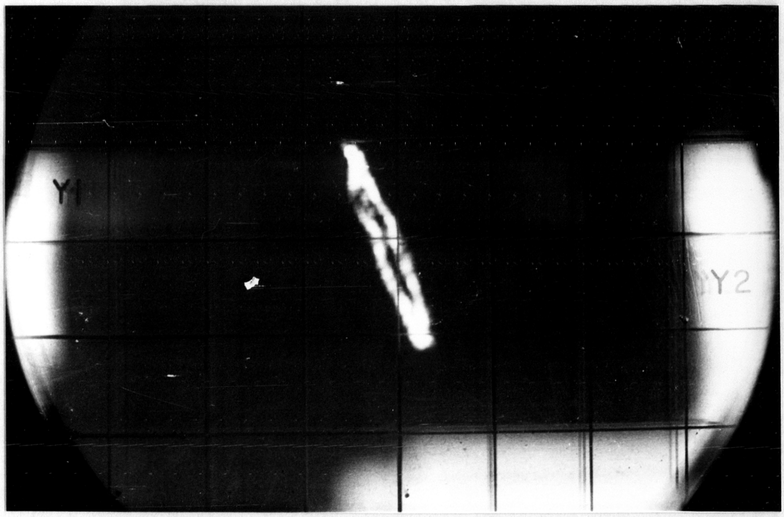


Fig. 27. Lissajous' figure, Δf = 0.5 MHz, r = 160 feet.

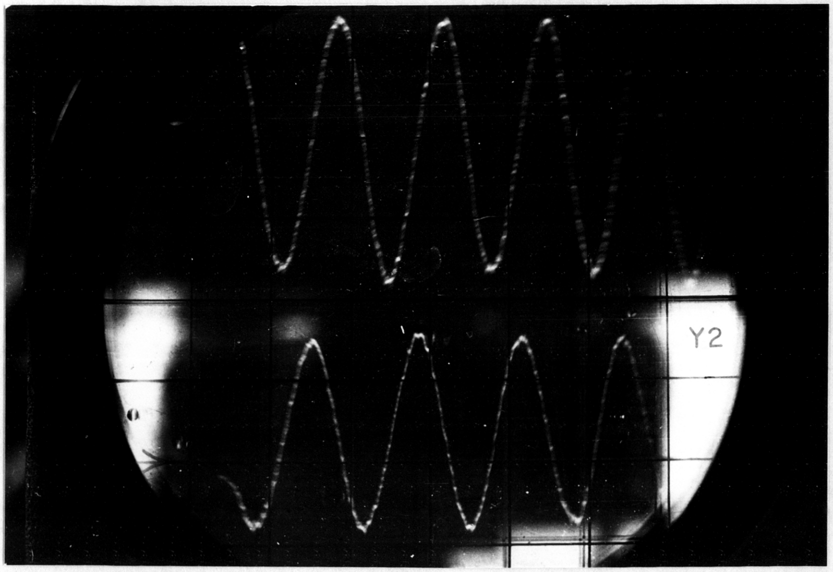


Fig. 28. Upper Trace: Doppler signal with frequency f_{D1}

Lower Trace: Doppler signal with frequency f_{D2}

(Δf = 1.5 MHz, r = 80 feet).

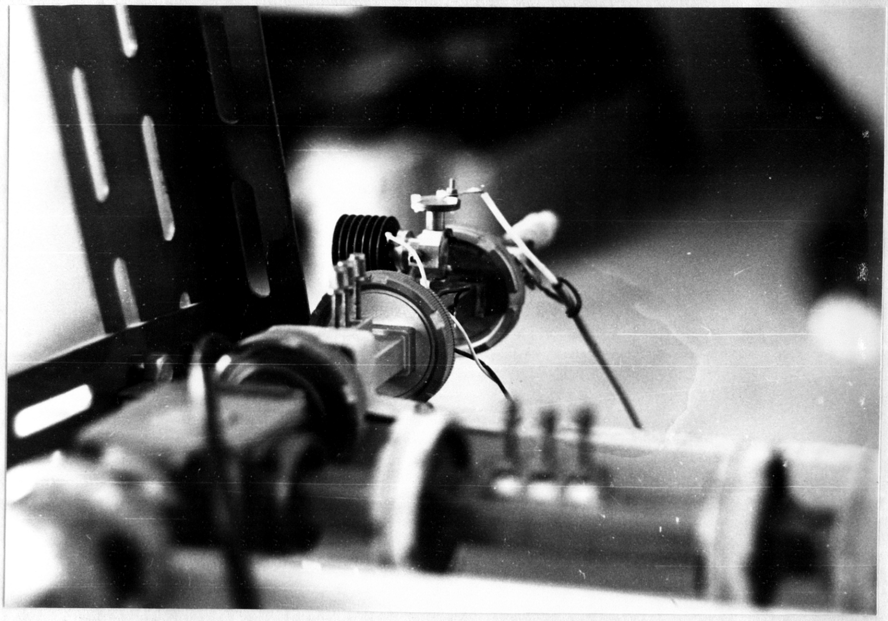


Fig. 29. Microwave Circuit. Showing the Gunn diode cavity with cooling fins. The tuning varactor diode is coupled to the cavity at the top.

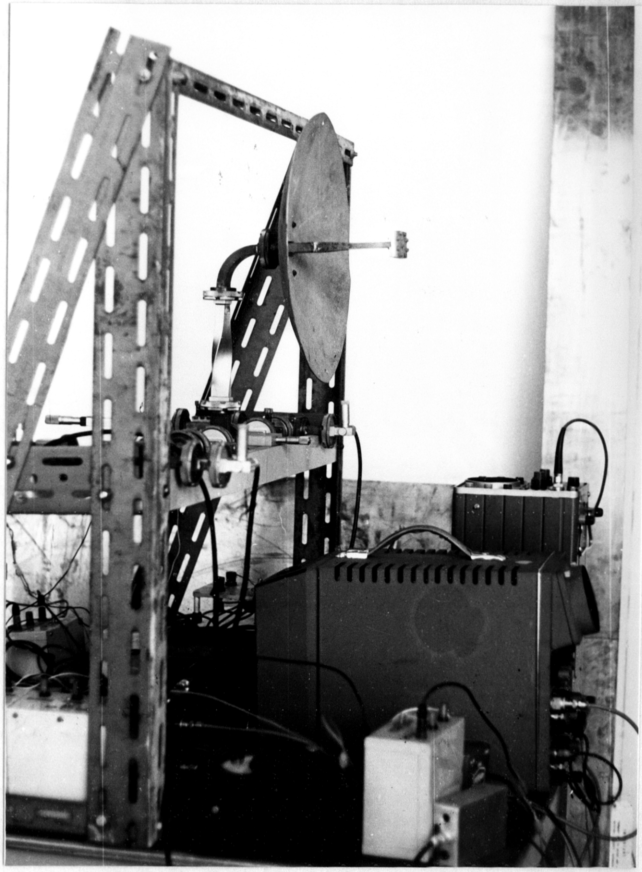


Fig. 30. General view of the equipment. The control circuit and Doppler channel amplifiers are below the microwave circuit. The video amplifier is in the right foreground.

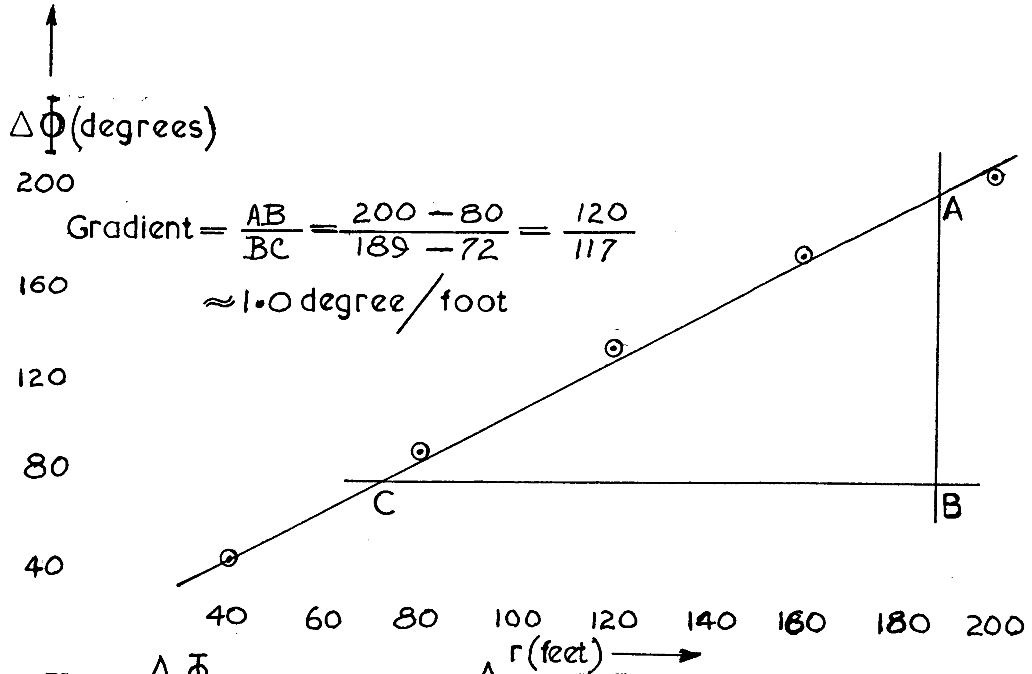


Fig. 31. $\Delta\Phi$ versus r, for Δf = 1.5 MHz

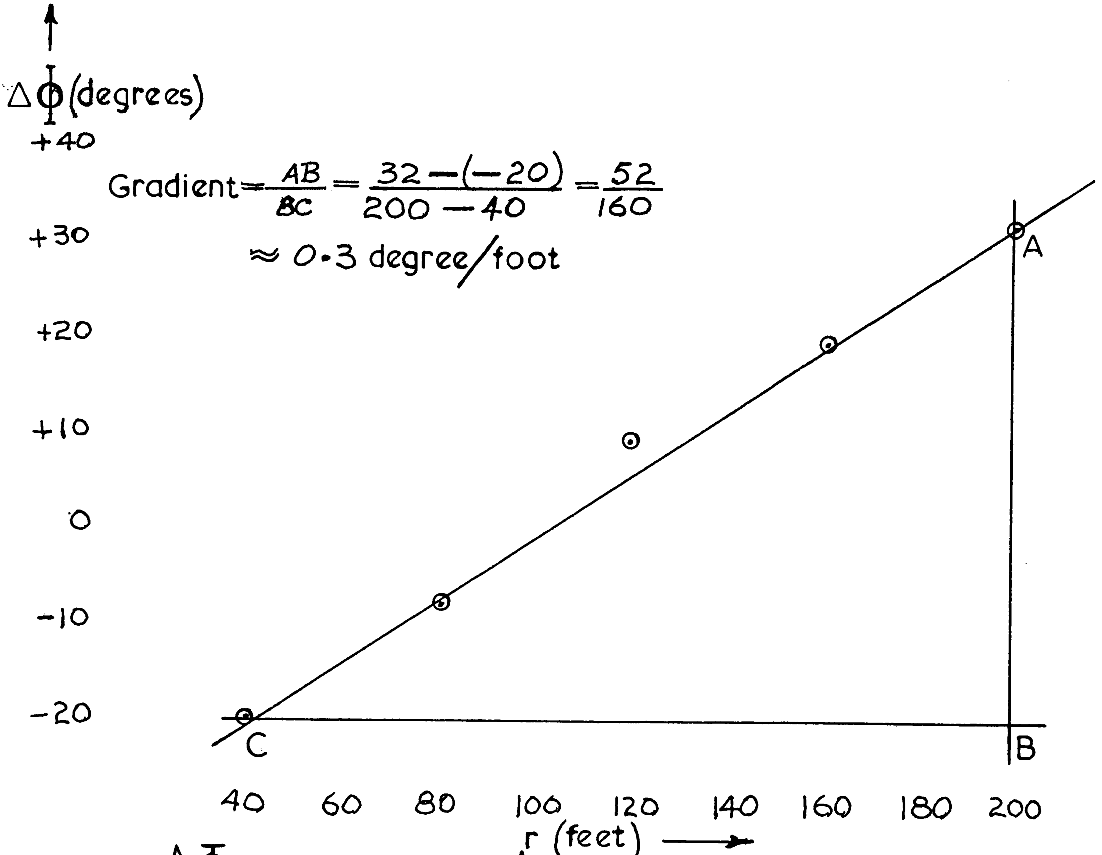


Fig. 32. $\Delta\Phi$ versus r, for Δf = 0.5 MHz

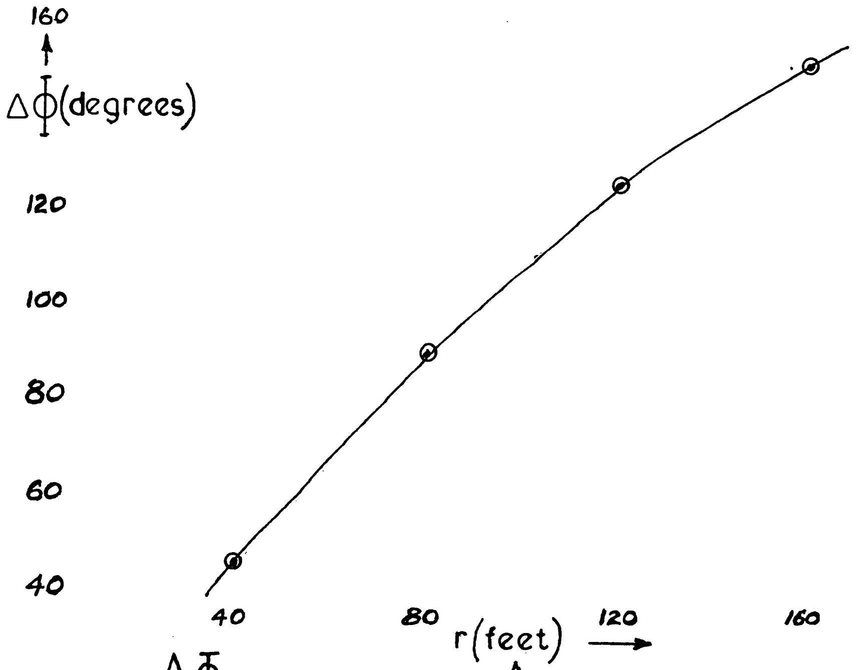


Fig. 33. $\Delta\Phi$ versus r, for Δf = 1.5 MHz, setting 1 of the 3-screw-tuner C (Fig.8).

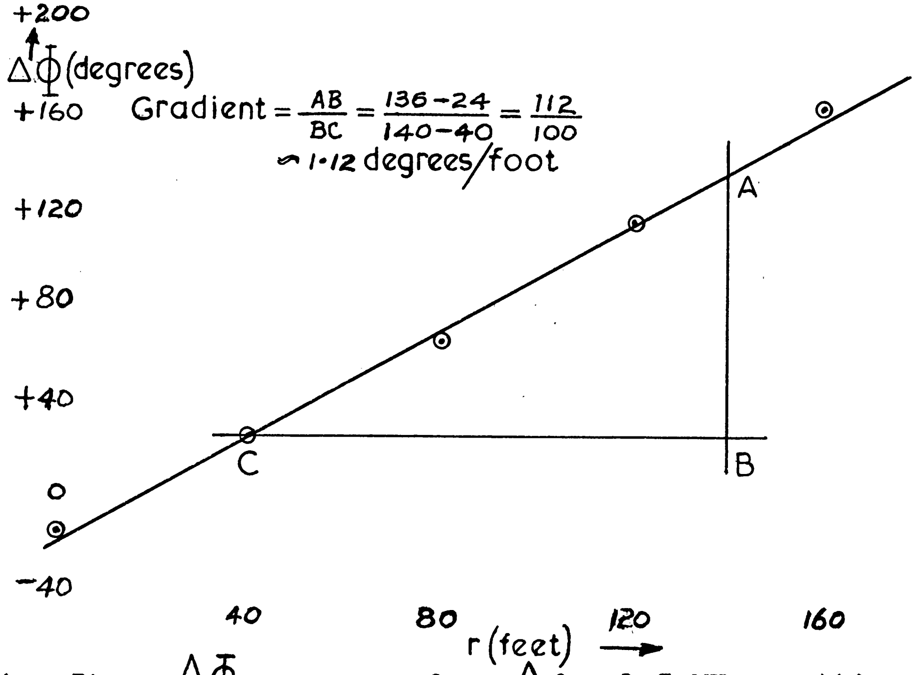


Fig. 34. $\Delta\Phi$ versus r, for Δf = 1.5 MHz, setting 2 of the 3-screw-tuner C (Fig.8).

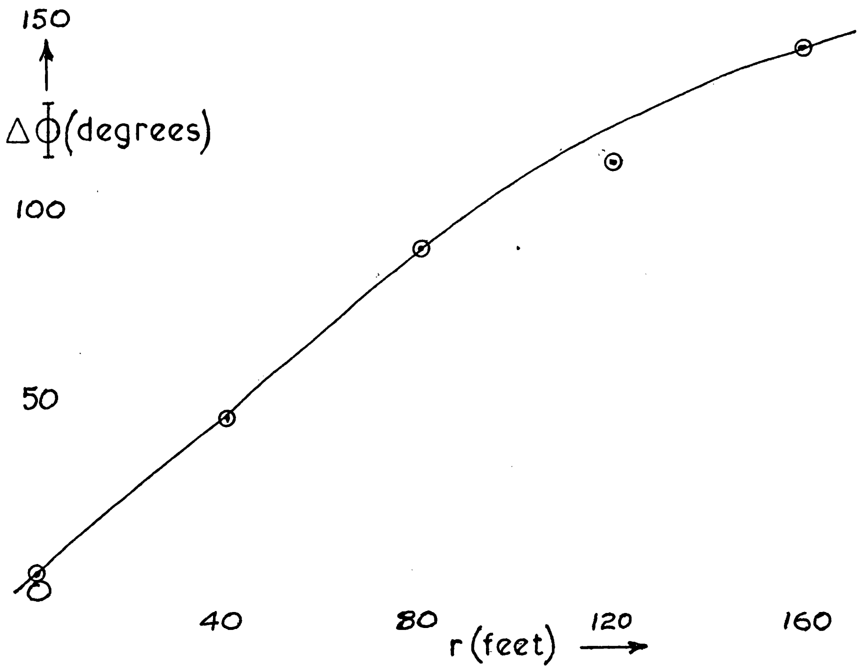


Fig. 35. $\Delta\Phi$ versus r, for Δf = 1.5 MHz, setting 3 of the 3-screw-tuner C (Fig.8).

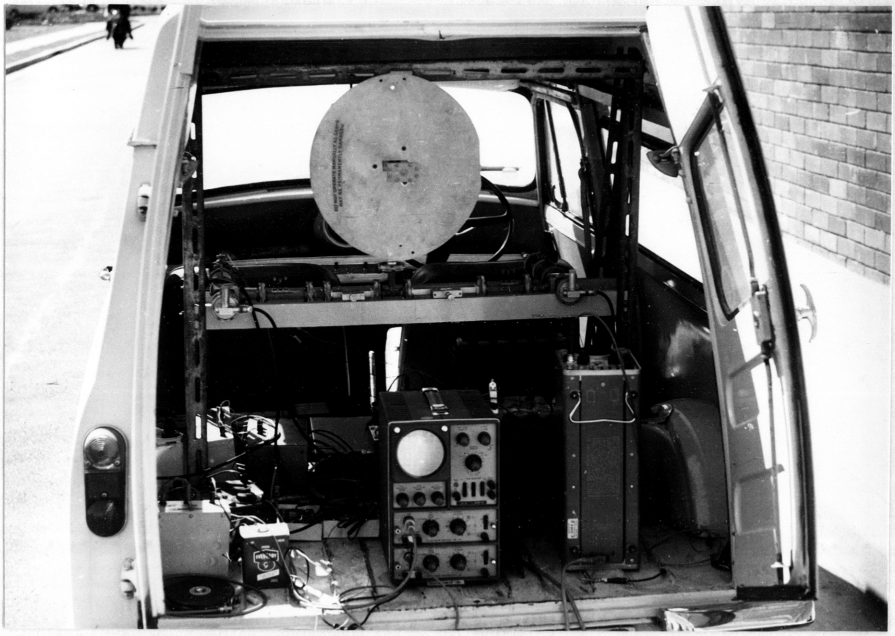
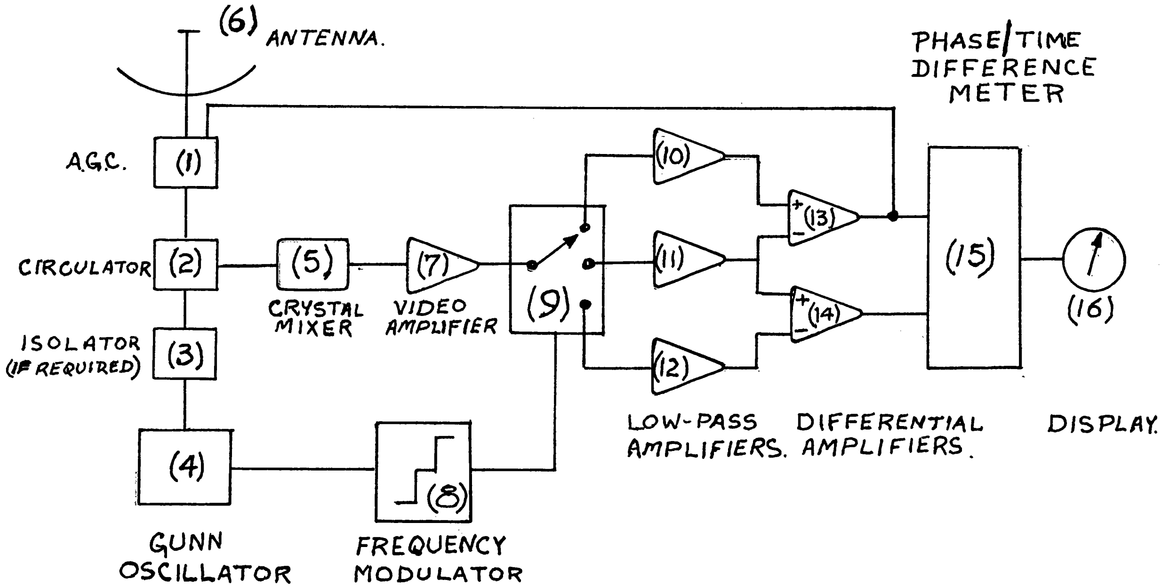


Fig. 36. Radar shown in the van. The Lissajous' figures were obtained on the oscilloscope facing the camera. The oscilloscope, with its axis vertical, to the right of the picture, was used to monitor the difference frequency.



(1) AGC (if required), for attenuation of strong local signals (BOYER[8]).

(7) VIDEO-AMPLIFIER, 1 MHz bandwidth.

(8) FREQUENCY MODULATOR, 3-step staircase waveform, 100 kHz repetition rate. Connected to the bias supply or to a varactor diode coupled to the cavity.

(9) SYNCHRONOUS SWITCH, driven by the frequency modulator.

(10), (11) and (12) LOW-PASS AMPLIFIERS, for filtering of the 100 kHz switching frequency.

(13) and (14) DIFFERENTIAL AMPLIFIERS, for cancellation of in-phase signals.

Fig. 37. Three frequency microwave spectrum CW Doppler radar, for simultaneous range measurement and clutter cancellation.